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ECONOMY
42218
200
*F=0010.00000kHz
FUNC:WAVE=SIN
JHD 162A
PCF8563
SQW
SCL
SDA
VCC
GND
ZS-042
A0 A1 A2

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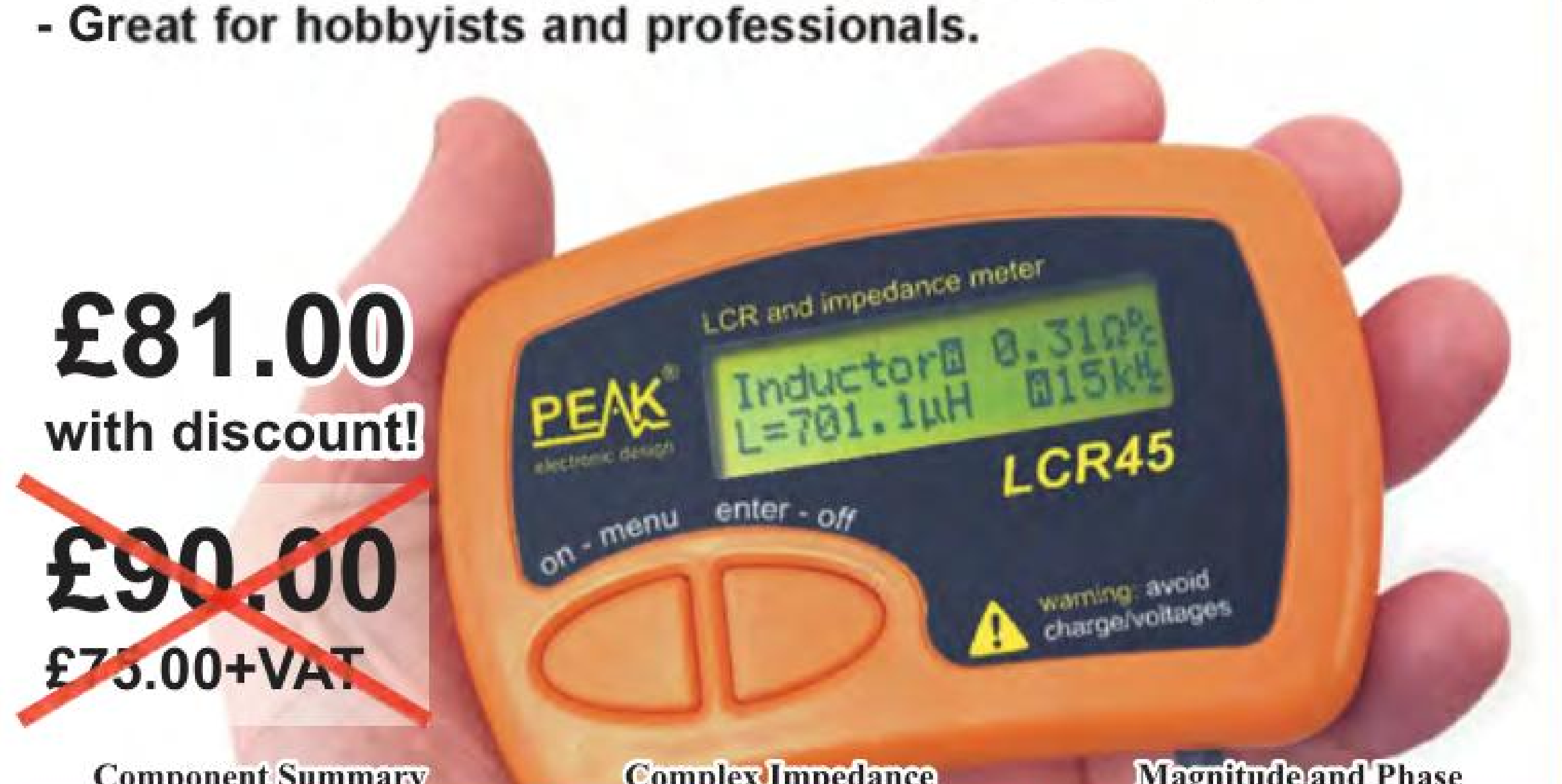
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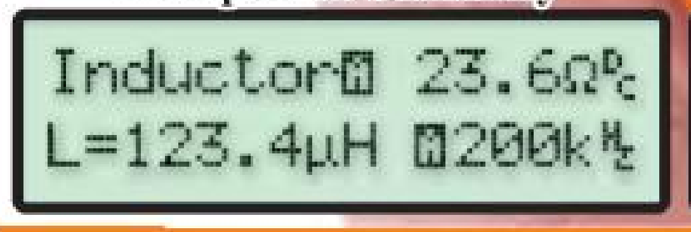
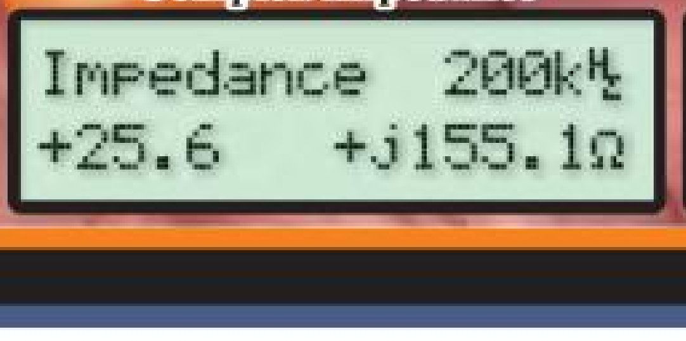


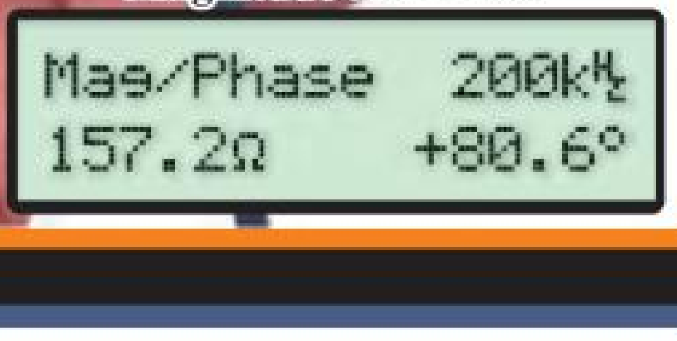
```
Inductor  23.6Ω
L=123.4µH  200kHz
```

Complex Impedance



```
Impedance   200kHz
+25.6    +j155.1Ω
```

Magnitude and Phase



```
Mag/Phase   200kHz
157.2Ω      +80.6°
```

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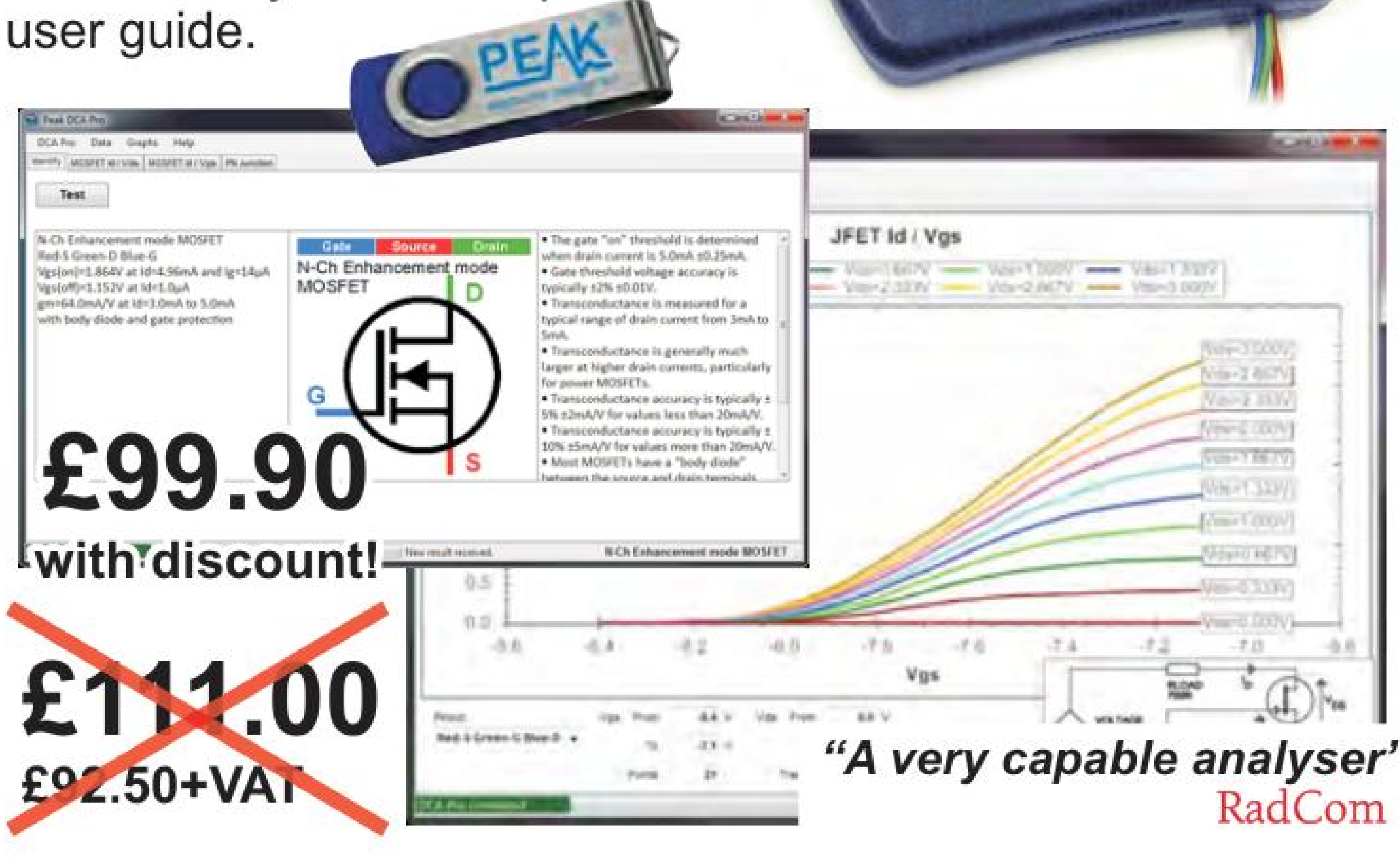
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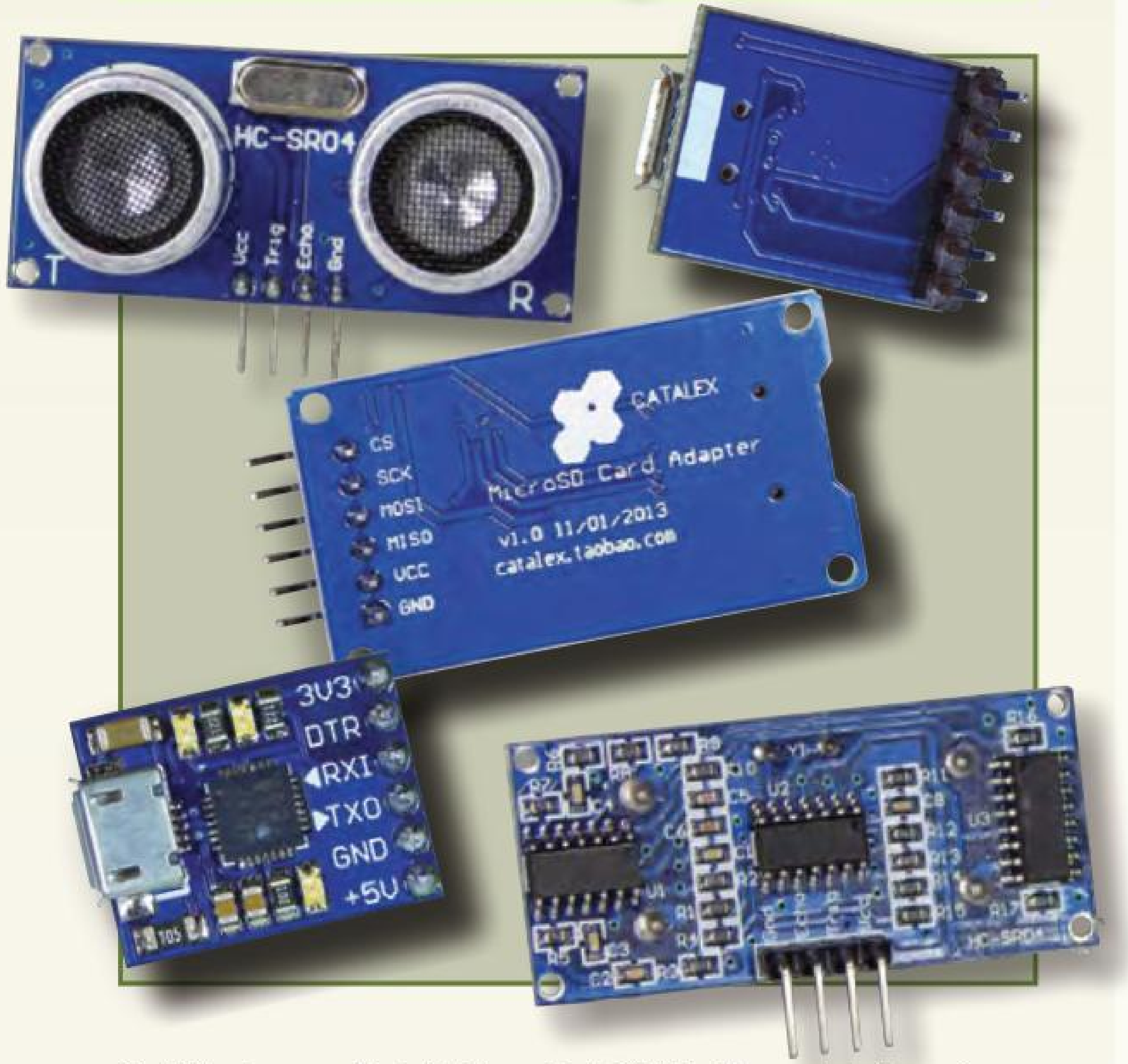

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Our April 2018 issue will be published on Thursday 1 March 2018, see page 72 for details.

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EVERYDAY PRACTICAL ELECTRONICS

Low-cost Asian modules

This month, we have a double serving of our new series looking at very cheap pre-built electronic modules coming from China. They offer exceptional value and a quick and (relatively) pain-free route to getting to grips with sensing and interfacing, and building a whole host of other electronic sub-systems that even a few years ago required considerably more expense and time to implement.

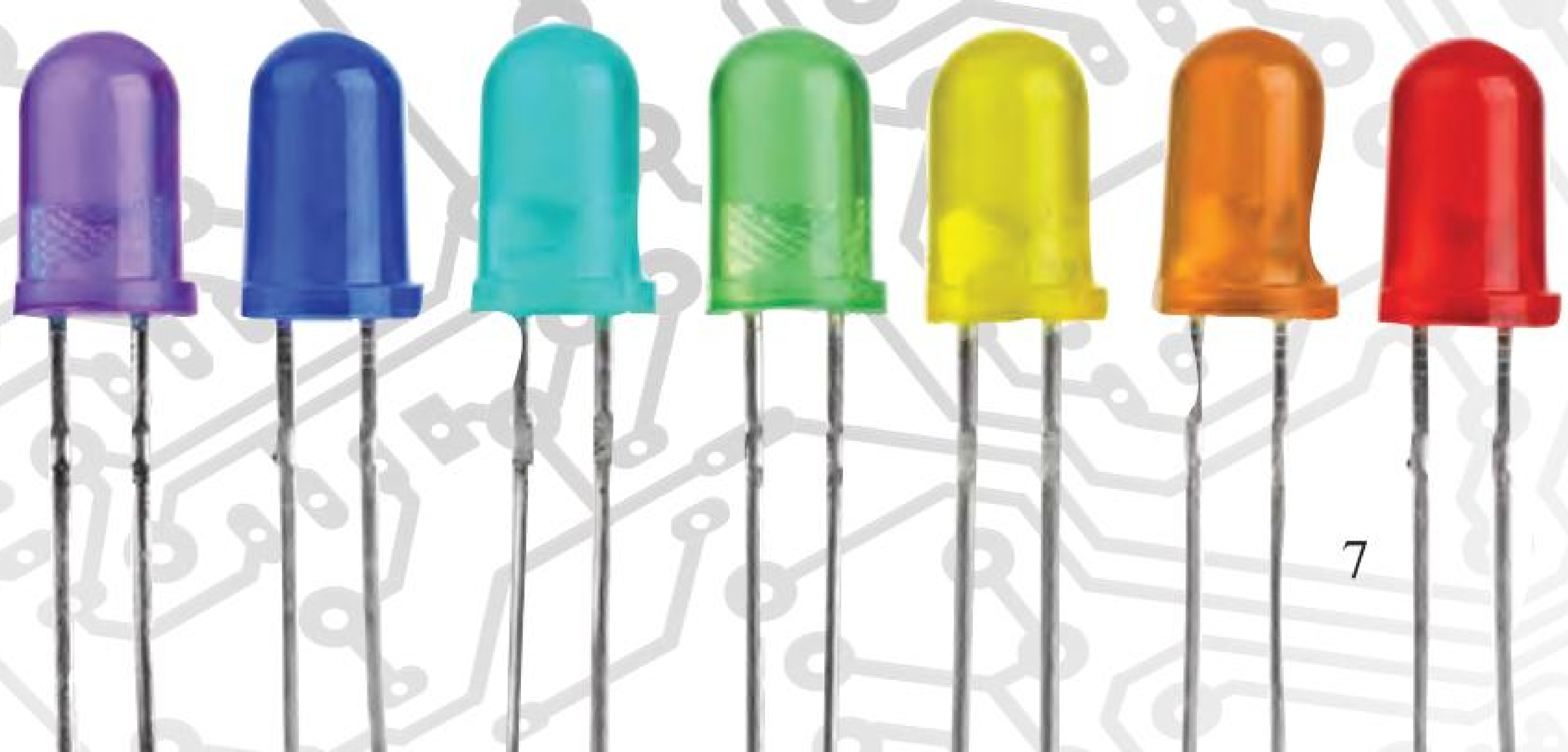
I have heard the criticism that such modules reduce electronics to little more than silicon Lego, but I strongly disagree. I'm sure similar arguments were levied against op amps when they first came out – why learn how to make a decent amplifier when a 741 can be bought for peanuts? The answer, of course, is that such 'Lego building blocks' complement basic skills, they do not replace them. They offer you a chance to build faster and to build more sophisticated designs, but as always, to get the most out of any raw material – whether it is a humble resistor, or a complete microcontroller system – understanding the basics is vital for getting good results. Use these cheap modules, enjoy them but make sure you keep your core skills sharp and up to date!

Back to the PIC n' Mix future

You will read at the end of this month's *PIC n' Mix* about Mike O'Keeffe's latest project – Chris and Ethan – his newborn twins. We send him and his wife Ciara our congratulations and best wishes for 2018. Quite understandably, Mike has asked for a few month's off from writing, and in April's issue I'm delighted to report that Mike Hibbett, our previous *PIC n' Mix* columnist will make a return. (Yes, you only get to write the *PIC n' Mix* column if you are called 'Mike'!)

I had a long discussion with Mike about what topic(s) he should choose and was really pleased that without any prompting he offered to write about a subject I have long wanted to see in *EPE* – the Fast Fourier Transform, thankfully 'FFT' for short.

The underlying mathematics behind FFT is quite sophisticated, certainly university level, and we won't concentrate too much on that, but what I do hope to achieve is a working design that enables readers to take advantage of this very useful algorithm. FFT is at the heart of digital spectrum analysers, digital filters and certain kinds of digital control systems. We won't be able to replace a £10k Agilent spectrum analyser, but we should be able to offer something that is genuinely useful – and fun!



NEWS

A roundup of the latest Everyday News from the world of electronics

Super cheap PCs and a visit from Mr Fox – report by *Barry Fox*

Build a PC from component parts and Microsoft will charge you £120 to download the Windows 10 Home Operating System to make it work (**http://bit.ly/2lmqNXV**).

'Free' PC

However, for the same price (or less) you can have a refurbished ex-office desktop with a legitimate copy of Windows 10 Home pre-installed, delivered to your door. I just did, for emergency use if my main PC is locked up doing upgrades, or has crashed… just when I might need it most. (See Micro Dream Ltd: **http://bit.ly/2lcq900**)

On the move on the cheap

What about for those of you on the move? If you need something light and it wouldn't be the end of the world if it were broken or stolen, then you can buy a brand new laptop with Windows 10 Home, for around £200. I borrowed a couple of them for a quick hands-on.

The Asus VivoBook E203 11.6-inch laptop weighs less than 1kg, is smaller than a piece of A4 paper and costs around £180. The spec is basic; Intel Celeron N3350 processor with 2GB of Ram. It's slow, of course, but has all the basic necessities – USB 3.0, HDMI, SD slot, Bluetooth, Wi-Fi and a webcam. I found the trackpad a bit sluggish, but a cheapo mouse easily fixes that.

Free PC – for the same price as a new Windows licence, Micro Dream Ltd will sell you a refurbished PC with a legitimate copy of Windows 10 Home pre-installed

The Windows 10 Home Acer Swift 1 has a 14-inch screen and rather higher spec; Intel Pentium Quad-core with 4GB RAM, but generally similar features – SD slot, Wi-Fi, USB 3.0 and HDMI. Battery life is better though. At £330 the price reflects this, but if I needed something to take where I wouldn't dare take my MacBook Pro, then the Acer would fit the bill nicely.

The real cost

But remember the old adage; it's not the cost of new hardware that matters as much as the pain of getting it up and running as a working tool. Even with no-cost open source software such as Open Office, Gimp graphics editing, VLC video player, Windows Defender security and the free edition of the excellent Macrium Reflect back-up (**www.macrium.com/reflectfree**) the time it takes to install and fine tune everything can sap the will to live.

With both laptops it took me many hours just to install all the security updates every Windows PC needs when first set up 'out of the box'. Just as you think you are finally done, the wretched thing is asking to install more updates, with more shut-downs, re-starts and dire warnings not to switch anything off and keep the mains power plugged in.

No wonder so many Windows PCs are inadequately protected – their owners will do almost anything to bypass the updates needed to keep them protected.

CCTV

Mr Fox visits Mr Fox – a night vision screen grab from Blink

I have previously tested several DIY CCTV systems (mainly for nature cams), and recommended Arlo from Netgear. Waterproof and wireless with night vision, the cameras use a home hub to send email alerts. But Arlo is expensive; originally around £300 for a two-camera system.

Now there is a similar wireless system from Blink. The XT is weatherproof, also wireless with night vision, and costs around £150. The XT is a very low power device, which should make battery life better than for Arlo. Both Arlo and Blink rely on Cloud storage and can be set to detect motion, record clips, send an alert, play the clip and offer live streaming.

The big difference is that Blink is controlled only by smart phone apps, with no option to work with PC. Also, whereas Arlo sends email alerts, Blink sends alerts only to the mobile app. But at a recent London launch for the XT, Peter Besen, Blink CEO, confirmed the company is working on PC control and viewing and email alerts. Predictably, Arlo's prices are now dropping towards parity.

Knock, knock

US company Ring sells the Video Doorbell, which shows who is at the door – even when the owner is miles away. As a visitor gets close, an IR camera sends their image off to the Cloud and rings a smartphone app that then displays the image. When the doorbell is pressed, the app uses VOIP to let the owner see and speak with whoever is at the door. Ring costs around £100.

Super cheap PC and a visit from Mr Fox – *continued*

The makers have now done a deal with Selfridges to sell a $100,000 version, with profits going to UK charities supporting the reformation of former criminals.

Each of the ten limited-edition Jewel Edition Video Doorbells is made of gold and encrusted with 2000 sapphires and 40 diamonds.

At the launch event in London, I suggested to the developer that I would be worried about leaving that much value outside on a doorstep, just asking to be stolen.

'Well, you will have a picture of the thief', he said; somewhat naively trusting in every ne'er-do-well to forget their balaclava!

Next-generation mobile technology for UK rail

The UK government has published its latest views on the implementation of the next iteration of mobile wireless communications technology: *Next Generation Mobile Technologies: An update to the 5G strategy for the UK*.

'5G' is the fifth generation of mobile communications technologies, which is designed to deliver a step change of ultrafast, low latency, reliable mobile connectivity that is able to support ever-larger data requirements, as well as wide-ranging new applications such as smart cities, autonomous vehicles and the Internet of Things.

The government has highlighted 2020 as the year when commercial rollout of 5G should begin in the UK, with 95% geographic coverage of the UK by 2022.

An interesting specific target set by the government is for high quality coverage on all major rail lines. The government has acknowledged what all rail commuters have known for years – 'relying on existing mobile operator coverage, even if supplemented by additional sites to infill poor coverage areas, is unlikely to deliver a sufficient level of connectivity to passengers on trains to meet future needs. This is partly because cuttings and the topography of the rail corridor mean that passengers are usually below the line of sight of masts, which are often primarily placed to provide coverage in populated areas, rather than to rail passengers … it is likely that trackside infrastructure will be required to deliver high quality, reliable coverage in areas of high passenger demand.'

The *Strategy* lists four key steps necessary for delivering high-quality rail data:

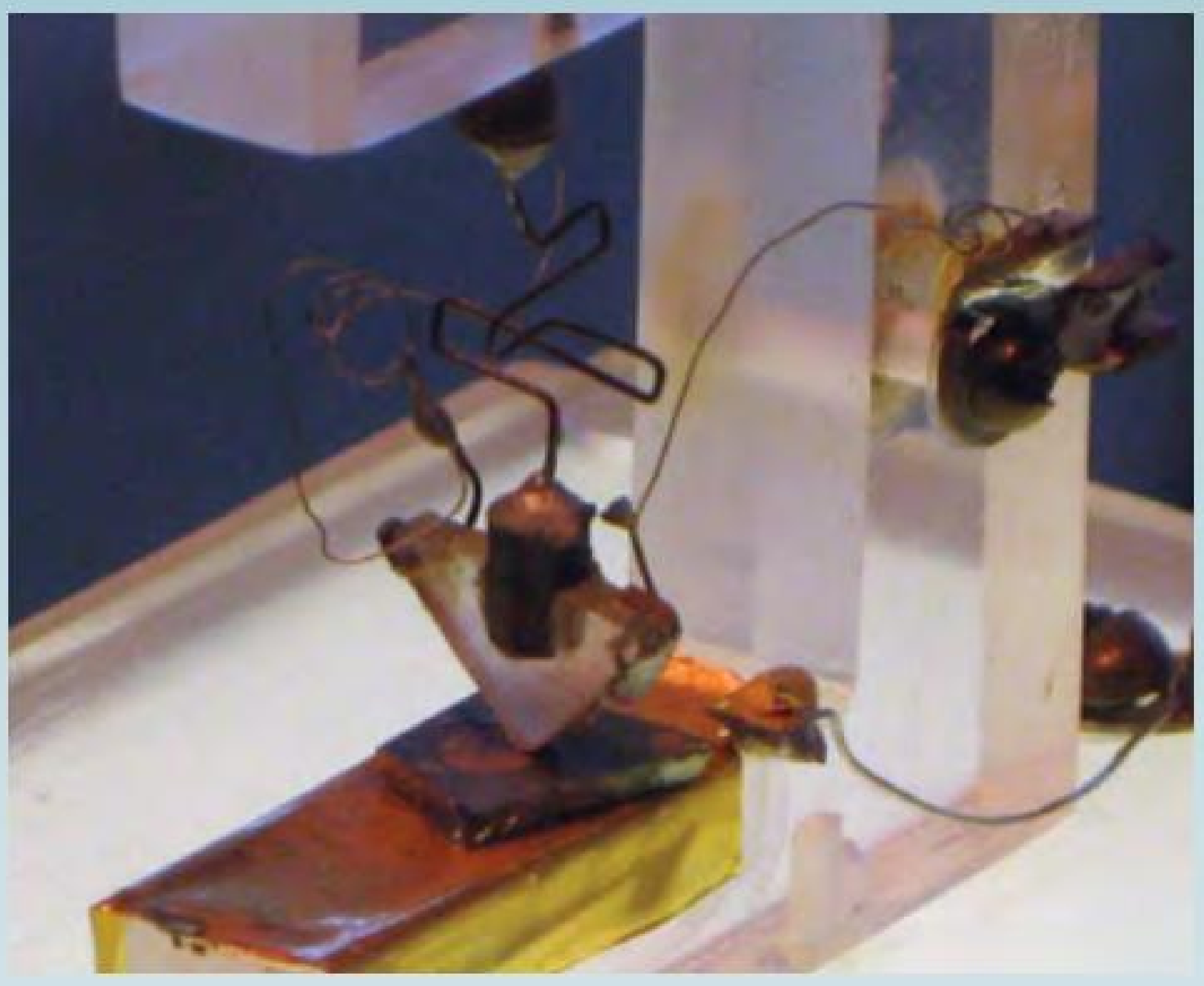
- Laying trackside fibre to provide the backbone of high-speed data
- Mounting wireless devices on masts and other infrastructure along the rail corridor
- Determining the radio access needs for delivering uninterrupted mobile connectivity, and ensuring the spectrum is made available
- Relaying signals into train carriages via on-board equipment

Trans Pennine initiative

To test potential solutions in a real-world environment, trials are planned for the Trans Pennine route. Trackside infrastructure will be deployed between Manchester and York, to enable the testing of options for track-to-train connectivity. They will begin by the end of 2018 and will help build a scalable model to deploy trackside infrastructure across the rail network.

The Trans Pennine route was chosen because of its challenging topography and the mix of passengers (commuters, day trippers, long/short-distance travellers), which will provide a good mix of real users to test demand.

Happy 70th birthday to the transistor!



The first transistor – the all-important semiconductor is a small (dark) slab of germanium immediately under the 'point' of the triangular contact.

Hard to believe, but just before Christmas 2017 the transistor turned 70.

The modern electronic era, based on solid-state semiconductor devices, started back in December 1947, when research scientists John Bardeen, William Shockley and Walter Brattain at Bell Laboratories announced the first working transistor. In their design, the researchers used germanium to create a 'point-contact' transistor.

Although the germanium point-contact design did work, it was expensive to make and prone to noise. These limitations meant that it was soon displaced by the now familiar bipolar 'sandwich' layout. Then, in 1954, germanium was replaced with silicon in the first commercial silicon transistor manufactured by Texas Instruments (TI). Just four years later, a TI researcher, Jack Kilby, produced the first integrated circuit.

Since those early days there have been many different types of transistor, but the fundamentals have been remarkably durable. This incredibly useful, scalable and flexible building block is now routinely manufactured by the billion in each microprocessor and high-capacity memory chip.

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Super-scary superflares

TechnoTalk

Mark Nelson

Climate change is the story you missed in 2017, argued one of Britain's quality newspapers at the end of last year. It seldom makes the headlines because so far its worst consequences have taken place thousands of miles away. Therefore, many folk assume it will never happen here, even though its effects make it an existential threat to life as we know it worldwide. But did you know we face another peril of equal and global dimensions that could destroy our electrical power grids, satellites, Internet and electronic devices?

GRUDGINGLY, MOST OF THE governments around the world have now accepted the reality of climate change. Last October, a project involving 24 academic institutions and intergovernmental organisations from across the globe stated that the effect of climate change on human health is now so severe that it should be considered 'the major threat of the 21st Century'. Major undoubtedly, but it's not the only one.

Frightening flare-up

According to two senior scientists, 'the most powerful superflares can serve as plausible drivers of extinction events' and 'the risk posed by superflares has not been sufficiently appreciated'. Is this sensationalism or wild exaggeration? I suspect the former, because these carefully moderated but absolutely authoritative statements were made by Harvard University professor Avi Loeb and postdoc candidate Manasvi Lingam. They have calculated with academic precision how much havoc and destruction a solar flare might wreak, the possible likelihood and timing of a superflare, our key vulnerabilities and what mitigation measures should be taken. Before we check out their findings it might help to explain the nature and origin of superflares.

A superflare is an exceptionally large example of a solar flare, which are sudden flashes of increased brightness of the sun. Solar flares generate a very broad spectrum of emissions, ejecting clouds of electrons, ions and atoms, along with electromagnetic radiation into outer space. It normally takes only a day or two for this ejecta to reach Earth.

Damaging effects

The greatest potential harm from these energetic bursts of ultraviolet radiation and high-energy charged particles would be the destruction of our ozone layer, causing DNA mutations and disrupting ecosystems. As Loeb explains, 'The sun is usually thought of as a friend and the source of life, but it could also be the opposite. It just depends on the circumstances.'

Global infrastructure is also at risk, as was demonstrated during the most significant solar flare in recent history. This was the so-called 'Solar Storm' or 'Carrington Event' that took place in 1859, when the flare was visible to a naked eye. It produced stunning red, green, and purple auroras down to tropical latitudes such as Cuba or Hawaii, which were of such brilliance that newspapers could be read easily in daylight. More seriously, the abnormally high levels of electromagnetic radiation induced dangerously high voltages in telegraph wires, rendering them inoperable. In North America, for instance, a telegraph operator was severely shocked when he accidentally touched an earth terminal causing an arc to strike between his forehead and the telegraph equipment. The next day, telegraphists discovered they could disconnect their batteries and still transmit messages using only the induced current from the sky.

Scientists scared

The Business Insider UK website states that if a Carrington Event happened today, the world would likely have to deal with the simultaneous loss of GPS, cellphone reception, and much of the power grid. Without satellite guidance aircraft fleets might have to be grounded globally. Unprotected electronic infrastructure could fail outright. According to Loeb, a modern Carrington Event could cause about $10 trillion of damage to power grids, satellites and communications. A flare just a bit stronger could even damage the ozone layer.

'Back then, there was not very much technology so the damage was not very significant, but if it happened in the modern world, the damage could be trillions of dollars. A flare like that today could shut down all the power grids, all the computers, all the cooling systems on nuclear reactors. A lot of things could go bad,' Loeb explained.

Serious risk?

Loeb and Lingam theorise that the next big solar flare will occur within the next 100 years or so, with a 12 per cent chance of it happening in the next decade. Asteroid impacts command all the attention when it comes to life-threatening space events, but after matching geological and solar observation data, the two scientists found that superflares would be just as deadly and are equally likely.

Mitigation measures might reduce the risk, but at a cost. Their proposed solution is a 'magnetic deflector' placed at a certain distance away from the Earth. Taking the form of a giant wire loop, its construction would make it a challenging project. Equally demanding would be devising an energy source to maintain the current flowing though the giant coil; although, in principle, it could be extracted from the sun by setting up photovoltaic panels in space.

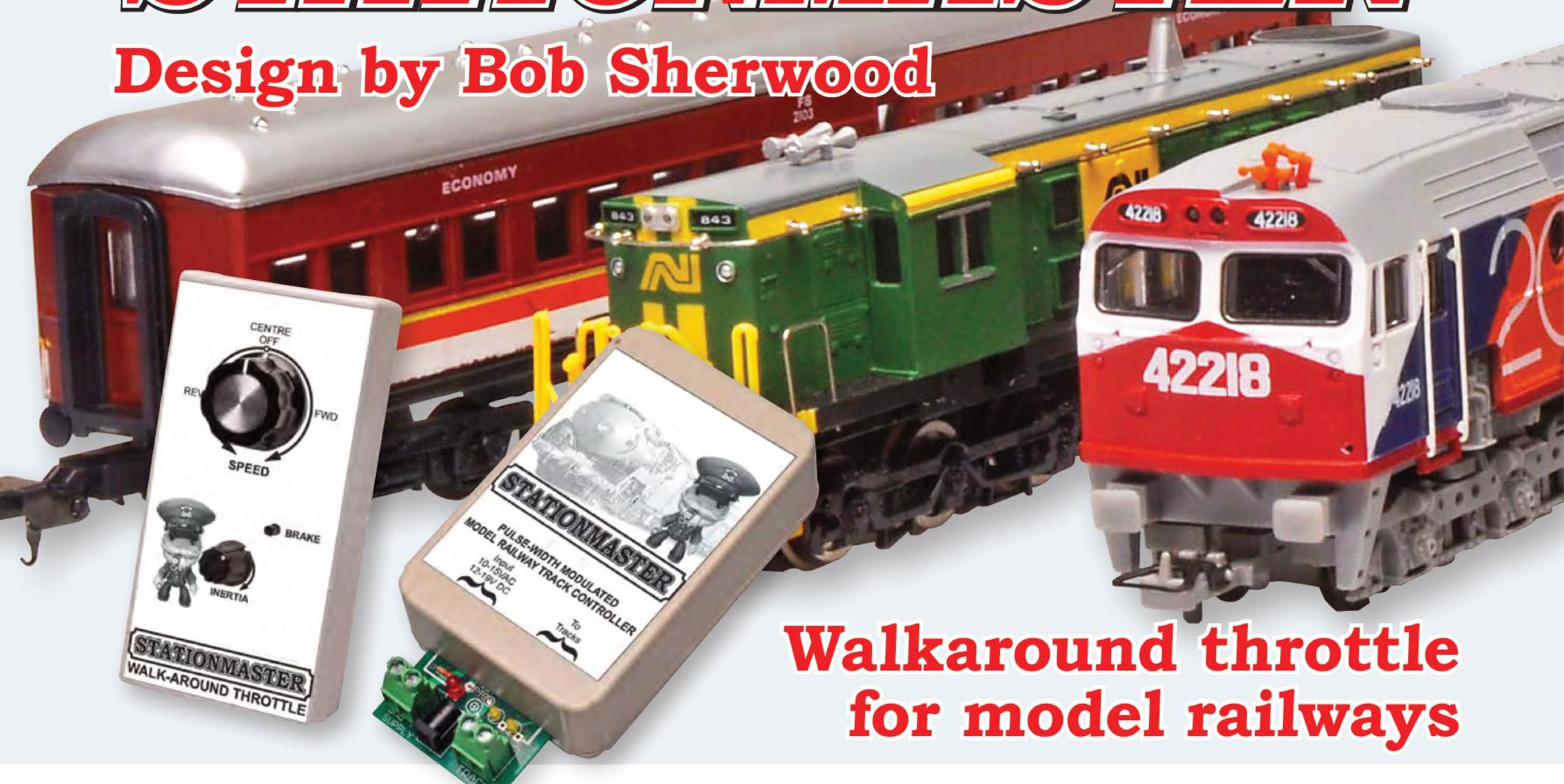
Reality checks

Loeb and Lingam say the total cost involved in lifting a 105-ton object into space would be around $100 billion, assuming that the payload cost per kg is $1000. This value is comparable to the total cost of the International Space Station, and is three to four orders of magnitude lower than the current world GDP or the economic damage from a flare around 100 years hence.

However, Gregory Laughlin, an astrophysics professor at Yale University, is not convinced. 'I'm not lying awake in bed at night worrying about solar superflares,' said Greg Laughlin from the Department of Astronomy at Yale University. 'But that doesn't mean that someone shouldn't be worrying about it,' he adds. 'I think that seriously diverting resources to build a wire loop in space would not be the best way to spend money. But thinking more about how solar superflares work and getting a sense of how our sun fits in with its peers would be a very valuable effort.'

STATIONMASTER

Design by Bob Sherwood



Walkaround throttle for model railways

Want to build a walk-around throttle for your model railway layout? This design is easy to build, yet provides useful features such as adjustable inertia, emergency braking and PWM control. It features a separate hand controller which you can plug into various sockets around your layout.

While Digital Command Control (DCC) is the bee's knees for large model railway layouts, a simple walk-around throttle is all you need for smaller layouts. And of course, there is nothing to stop you using this controller on a large layout, as well.

The benefit of a speed controller with a hand-held walk-around controller is that you can plug it into sockets at various points around your layout. This *Stationmaster* design by Bob Sherwood uses cheap, readily available Telecom-style RJ sockets and plugs. Your layout can have one socket or many, depending on how many you want, and you can use standard flat or curly leads.

Chances are you already have a spare AC or DC power supply that would be suitable to run the *Stationmaster*. Anything from 12V DC or 10VAC at 1A up to 25V DC or 18VAC at 5A would do the job. 1A will be plenty for a single locomotive, but if you're planning to run several on the same tracks, you will need at least two or three amps.

If you already have a train controller but it's a variable DC output type, you will want to upgrade to the *Stationmaster* because as you have probably noticed, any time the locomotive hits a dirty section of track at a low DC voltage, it tends to slow down, lurch or even stop. That's much less of an issue with PWM (pulse width modulation) drive because you will be applying higher peak voltages to the track.

The PWM voltage is applied to the track by an H-bridge IC. The operation of an H-bridge is shown in Fig.1, and four possible switch conditions are shown. Here we are showing the H-bridge as comprising four switches, although in the *Stationmaster* they are of course N-channel MOSFETs.

Fig.1(a) shows the default state with all switches off. In this state the motor is not connected to anything, and so if the locomotive is moving, it will continue to move but will slow down naturally due to friction in the wheels, gearing and motor. If the locomotive is not moving, it would be possible to push it along the track and it may roll down a steep grade on its own.

In Fig.1(b), switches S1a and S2b are closed. One end of the motor is connected to the positive supply and the other end to ground, so the motor is driven in one direction. In Fig.1(c), the opposite pair of switches is closed, and so the motor drive polarity is reversed and the motor will rotate in the opposite direction.

Features and specifications

- Walkaround hand controller
- Controls: forward/reverse, speed, inertia (momentum), emergency brake
- Indicators: power on, forward/reverse drive, track voltage indicators
- Short-circuit protection
- Output current: up to 3.5A; adjustable current limit
- Supply voltage: 12-25V DC, 10-18VAC
- Quiescent current: 20mA
- PWM frequency: ~8kHz

In Fig.1(d), switches S1b and S2b are closed, and so the motor is effectively shorted out. This will provide significant braking. If the locomotive is moving, it will quickly come to a halt and if it is stationary, it will be difficult to move and will not roll down a steep grade. If the opposite set of switches were closed (ie, S1a and S2a), the effect would be the same.

All four switches plus the control logic and gate drive circuitry in the *Stationmaster* are integrated into a single IC, a Texas Instruments DRV8871 H-bridge. One important feature of this IC is that it contains protection logic to prevent the wrong pair of switches from being closed, resulting in the power supply being shorted out.

Speed control is achieved by switching rapidly between the configuration of Fig.1(a) and either of Fig.1(b) or Fig.1(c), depending on the direction of travel.

The more time the H-bridge spends in the state of Fig.1(a), the lower the locomotive speed. With a PWM control scheme, the rate at which the H-bridge alternates between these configurations is fixed and speed is controlled by how much time it spends in the two states. The percentage of the time where voltage is applied to the tracks is known as the duty cycle; a higher duty cycle results in a higher speed.

Circuit description

The complete *Stationmaster* circuit is shown in Fig.2; it consists of two main sections. On the left is the PWM waveform generation circuitry and on the right, the DRV8871 H-bridge IC and associated components, to provide the high-current drive to the locomotive tracks.

The PWM-generation circuitry is based on IC1, a TL084 and IC2, an MC14584 hex schmitt trigger inverter. Two of the op amp stages, IC1a and IC2b, combine to form an ~8kHz triangle-wave generator. IC1b is configured as an integrator, with its pin 5 non-inverting input connected to a 2.5V half-supply rail derived from the 5V rail via two 220Ω resistors and a 1µF filter capacitor.

When its pin 6 inverting input is above 2.5V, the output voltage at pin 7 drops at a constant rate, whereas when the pin 6 input is below 2.5V, the output voltage at pin 7 rises at the same rate. Op amp stage IC1a is configured as a comparator with hysteresis, and its output is low when its pin 3 input is below 2.5V and high when its input is above 2.5V.

This input is fed via a divider from the output of IC1b, with the other end of the divider connected to its pin 1 output. So essentially, this completes the feedback path causing IC1b to oscillate, as well as defining the amplitude of the triangle wave it produces, by the ratio of the 1kΩ and 3.3kΩ resistors.

When output pin 1 of IC1a is low, at say 0.9V, output pin 7 of IC1b will need to rise above 3V in order to switch the output of IC1a high. You can confirm this by calculating the voltage at pin 3 (in the middle of the divider): ((3V × 3.3kΩ) + (0.9V × 1kΩ)) ÷ (3.3kΩ + 1kΩ) = 2.51V. Similarly, when output pin 1 of IC1a is high, at say 4.05V, output pin 7 of IC1b will need to fall below 2V in order to switch the output of IC1a low; ((2V × 3.3kΩ) + (4.05V × 1kΩ)) ÷ (3.3kΩ + 1kΩ) = 2.48V.

So these will be the approximate maximum and minimum voltages of the triangular waveform at output pin 7 of IC1b, with a maximum of around 3V, a minimum of around 2V and thus a peak-to-peak voltage of around 1V.

The actual waveforms produced by the prototype are shown in the oscilloscope grab of Fig.3. The waveform at pin 1 of IC1a is the green trace, while that at pin 7 of IC1b is the blue trace. As you can see from the measurements at the bottom of the screen, the actual peak-to-peak voltage of the triangle wave is 880mV and the frequency is 9.43kHz (the actual frequency will vary depending on circuit tolerances, but it is not critical).

The triangular wave is converted into a variable-duty-cycle PWM signal by comparing its amplitude to that of a DC control signal, which varies somewhere between its minimum and maximum voltages. The higher the control signal voltage, the higher the PWM duty cycle. However, the situation is complicated by the fact that we need to be able to drive the locomotive in either direction and that we also need a 'dead band' when the speed pot is set somewhere around the middle, where there is no drive at all.

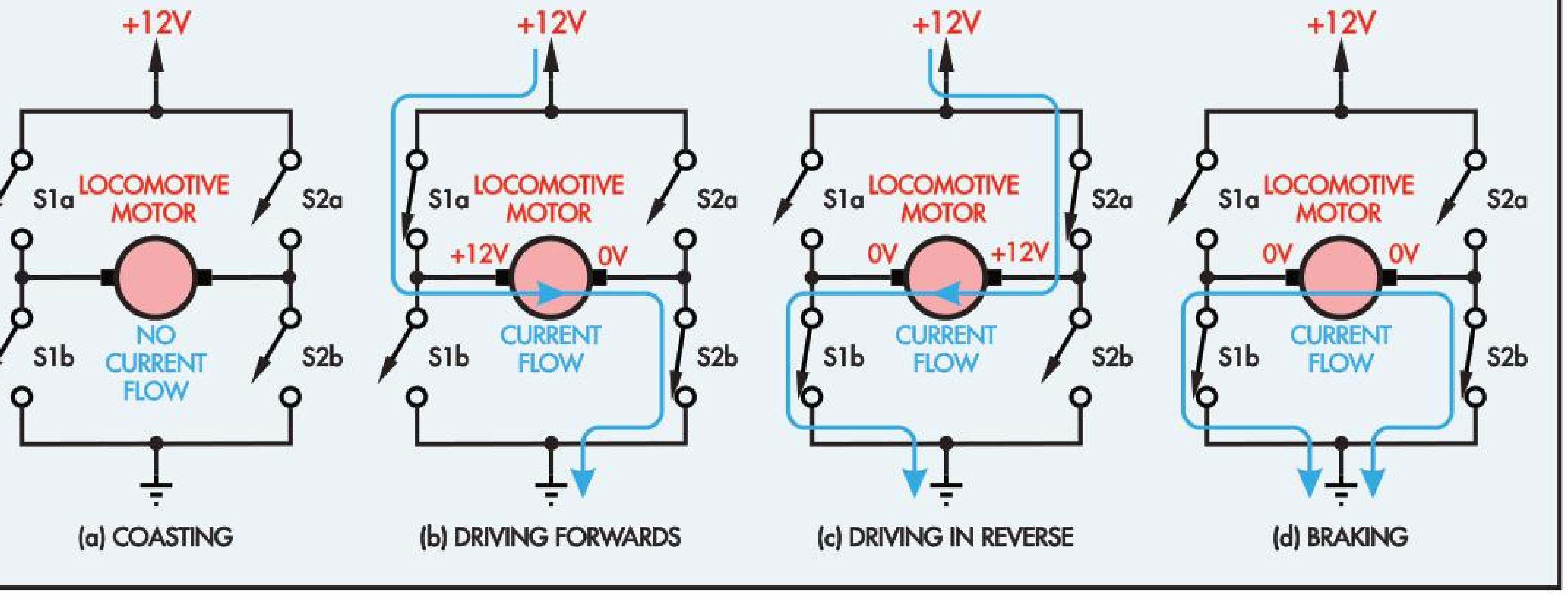
This situation is handled by using two comparators along with two triangle waveforms that have slightly different DC levels. The other two stages of op amp IC1 – ie, IC1c and IC1d – are used for these comparators, and the waveform from the pin 7 output of IC1b is coupled to two of their inputs (pins 9 and 12) via 100nF capacitors. The DC bias for these two pins is provided by a resistor network across the 5V supply comprising two 47kΩ fixed resistors, an 18kΩ resistor and 20kΩ trimpot (VR1) which is connected as a rheostat (ie, variable resistor).

Thus, input pin 9 of IC1c has a DC level between 2.84V and 3.22V, while input pin 12 of IC1d has a DC level between 1.78V and 2.16V, depending on the setting of VR1. The average of these two voltages will be very close to the 2.5V half-supply rail. The further apart these two voltages are, the larger the 'dead band' will be, allowing the speed control potentiometer to be rotated over a larger part of its range without any drive to the locomotive.

This adjustment is necessary to allow for variations in the amplitude of the triangle waveform; VR1 is adjusted until the waveforms no longer overlap, so that there is no drive to the locomotive tracks with the speed pot in its central position.

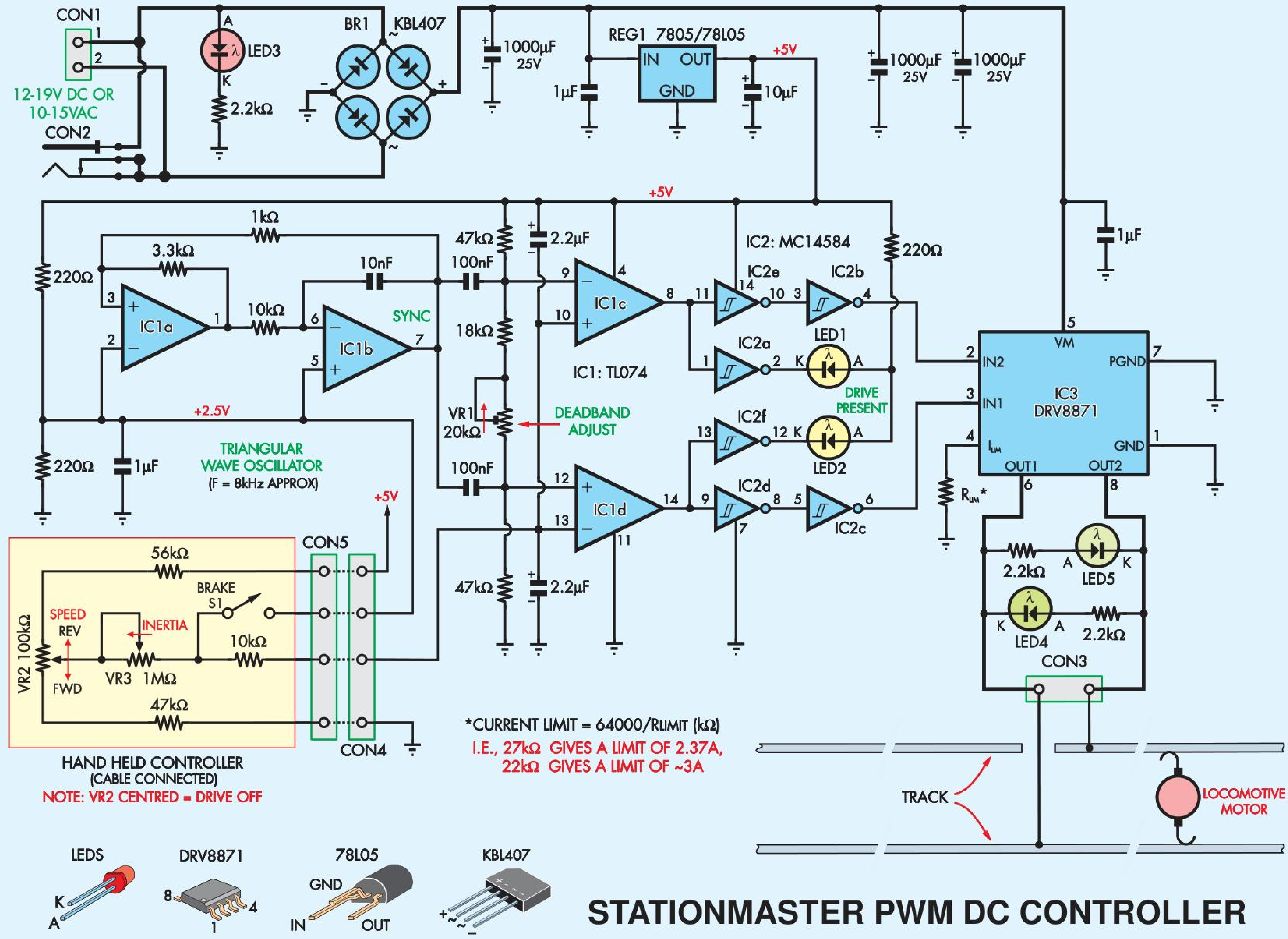
Also, there's no guarantee that when its speed pot is in its half-way position, it will necessarily be at exactly half its nominal resistance value. Indeed, if using a pot with a central detent, it would be very annoying if the loco slowly moved in one direction or the other. So the dead band needs to be so that the loco tracks get no drive with the speed pot at its half-way point.

The two DC-biased triangle waveforms can also be seen in Fig.3, with pin 9 of IC1c in yellow and pin 12 of



(a) COASTING (b) DRIVING FORWARDS (c) DRIVING IN REVERSE (d) BRAKING

Fig.1: four of the five possible configurations of an H-bridge (the fifth is not used in our application). The voltage across the motor and the current flow path is shown, assuming a nominal 12V DC supply. In case (d), the current flow direction depends on the direction of motor rotation at the time of braking. The switches are usually discrete MOSFETs (they may also be internal to an IC) as in the *Stationmaster*.


CON1
12-19V DC OR
10-15VAC
CON2
A
LED3
K
2.2kΩ
BR1
KBL407
1000µF
25V
REG1 7805/78L05
IN
OUT
GND
1µF
+5V
10µF
1000µF
25V
1000µF
25V
+5V
1kΩ
3.3kΩ
220Ω
IC1a
10kΩ
10nF
IC1b
SYNC
100nF
47kΩ
2.2µF
18kΩ
IC1c
IC1: TL074
IC2: MC14584
IC2e
IC2b
IC2a
LED1
DRIVE
PRESENT
IC2f
LED2
IC2d
IC2c
220Ω
1µF
IN2
IN1
VM
PGND
IC3
DRV8871
GND
OUT1
OUT2
I_{LIM}
R_{LIM}*
+2.5V
VR1
20kΩ
DEADBAND
ADJUST
220Ω
1µF
TRIANGULAR
WAVE OSCILLATOR
(F = 8kHz APPROX)
100nF
IC1d
47kΩ
2.2µF
+5V
56kΩ
BRAKE
S1
SPEED
REV
INERTIA
VR2 100kΩ
FWD
VR3 1MΩ
10kΩ
47kΩ
CON5
CON4
HAND HELD CONTROLLER
(CABLE CONNECTED)
NOTE: VR2 CENTRED = DRIVE OFF
*CURRENT LIMIT = 64000/RLIMIT (kΩ)
I.E., 27kΩ GIVES A LIMIT OF 2.37A,
22kΩ GIVES A LIMIT OF ~3A
2.2kΩ
LED5
LED4
2.2kΩ
CON3
TRACK
LOCOMOTIVE
MOTOR
LEDS
DRV8871
78L05
KBL407
GND
IN
OUT

STATIONMASTER PWM DC CONTROLLER

Fig.2: the complete circuit diagram for the *Stationmaster*, with the hand controller circuitry shown in the box at lower left. IC1a and IC1b generate a triangle waveform at around 8kHz, and IC1c and IC1d compare this to the control signal from speed pot VR2. The outputs of IC1c and IC1d are PWM signals which are squared up by schmitt trigger inverter IC2 and fed to H-bridge IC3 to drive the tracks.

IC1d in mauve. As you can see, VR1 has been adjusted so that the minimum voltage of pin 9 is above the maximum voltage of pin 12.

Speed, inertia and brake controls

The speed, inertia and brake controls consist of two pots and a momentary switch, and are usually mounted in the separate hand controller unit which is attached to the main board by a telephone cable.

Normally, a two-metre cable is about right; however, you can use a longer or shorter cable if necessary. There are provisions to mount these controls inside the main unit; however, we won't go into details about that option since we think most people will want to use the hand controller for walk-around operation.

The controls are shown at lower left in the circuit of Fig.2. Speed control pot VR2 is effectively connected across the 5V supply with padding resistors at either end to limit the voltage at its wiper so that it varies over an appropriate range to go from full speed in the forward direction to full speed in reverse, without too much of a dead zone at either end.

The inertia potentiometer is wired as a rheostat (variable resistor) and is in series with the return signal from the speed pot's wiper. The other end of the inertia pot is fed to a pair of 2.2µF capacitors on the main board, via a 10kΩ fixed resistor, so the higher a resistance the inertia pot is set to, the more slowly the voltage across these 2.2µF capacitors change. This simulates a locomotive with more inertia (mass), so its speed will change more slowly when the speed pot is rotated.

Brake switch S1 bypasses both the speed and inertia pots and connects the 2.5V mid-rail supply directly to the 10kΩ capacitor, which rapidly charges/discharges the 2.2µF capacitors on the main board until the locomotive has stopped and it will remain stopped until the brake switch is released; if the speed pot is at its midpoint after the brake is released, the loco will not move off again.

Note that braking is not instant, as this may cause the locomotive(s) to derail, but it will stop the loco(s) significantly faster than simply winding the speed pot back to its central position.

Track drive

The output of op amp (comparator) IC1c goes high when the speed control signal at its pin 10 non-inverting input is above the triangle waveform at its pin 9 inverting input. Thus, its output duty cycle increases with clockwise rotation of the speed pot.

Similarly, the output of op amp (comparator) IC1d goes high when the speed control signal at its pin 13 inverting input is lower than the triangle waveform at its pin 12 non-inverting input. Thus, its output duty cycle increases with anti-clockwise rotation of the speed pot.

As stated earlier, VR1 is adjusted so that the output of both comparators remain constantly low with the speed pot at its halfway point. This condition is shown in the scope grab

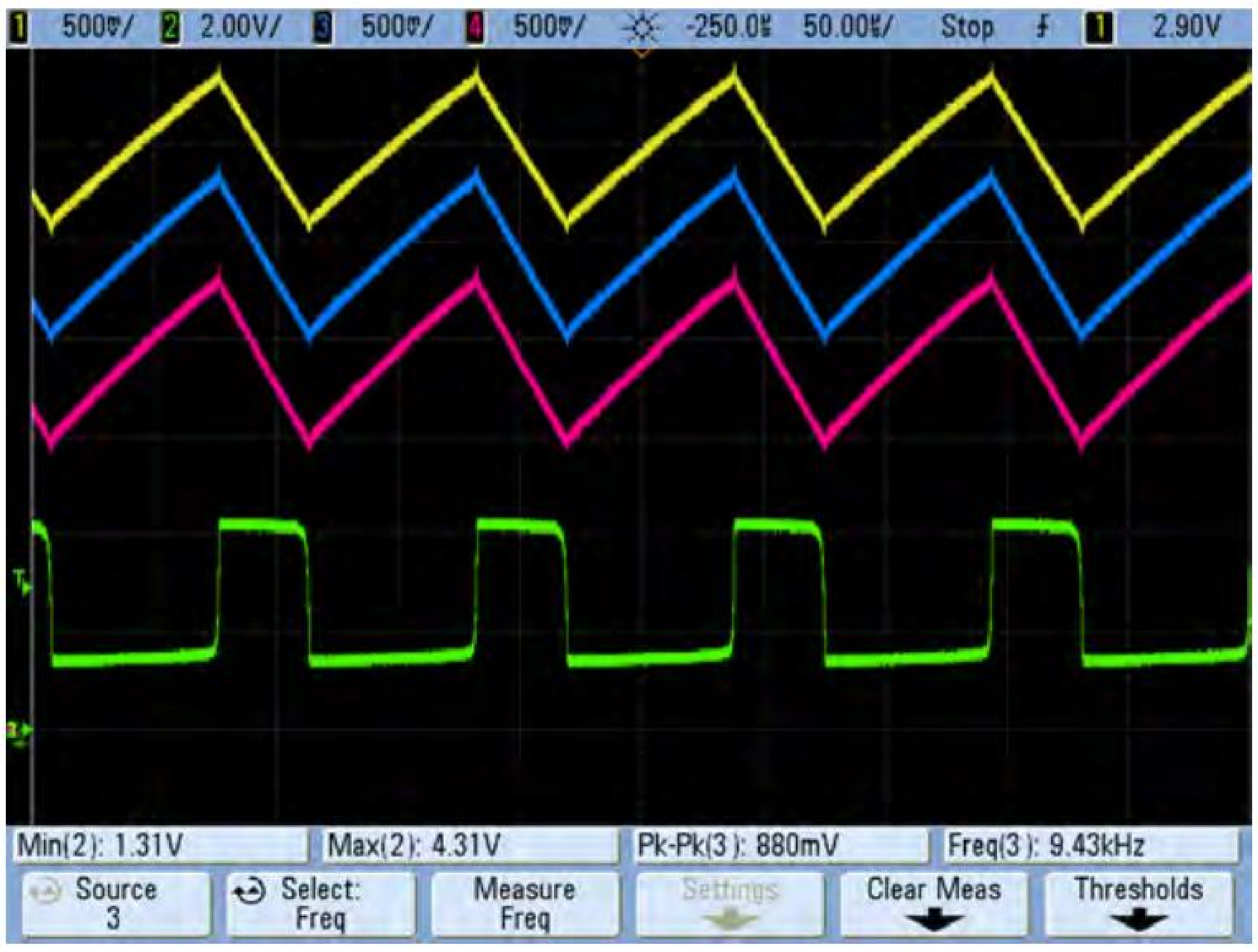


Fig.3 the blue trace is the triangle waveform at pin 7 of IC1b. It has a frequency of 9.43kHz and an amplitude of 880mV peak-to-peak. The yellow and mauve traces are the DC-shifted versions of this waveform at pins 10 and 13 of IC1 respectively. The green trace shows the pulse applied to pin 6 of IC1b which is in-phase with the triangle waveform and has a maximum voltage of 4.31V and minimum of 1.31V, limited by the drive capability of the op amp.

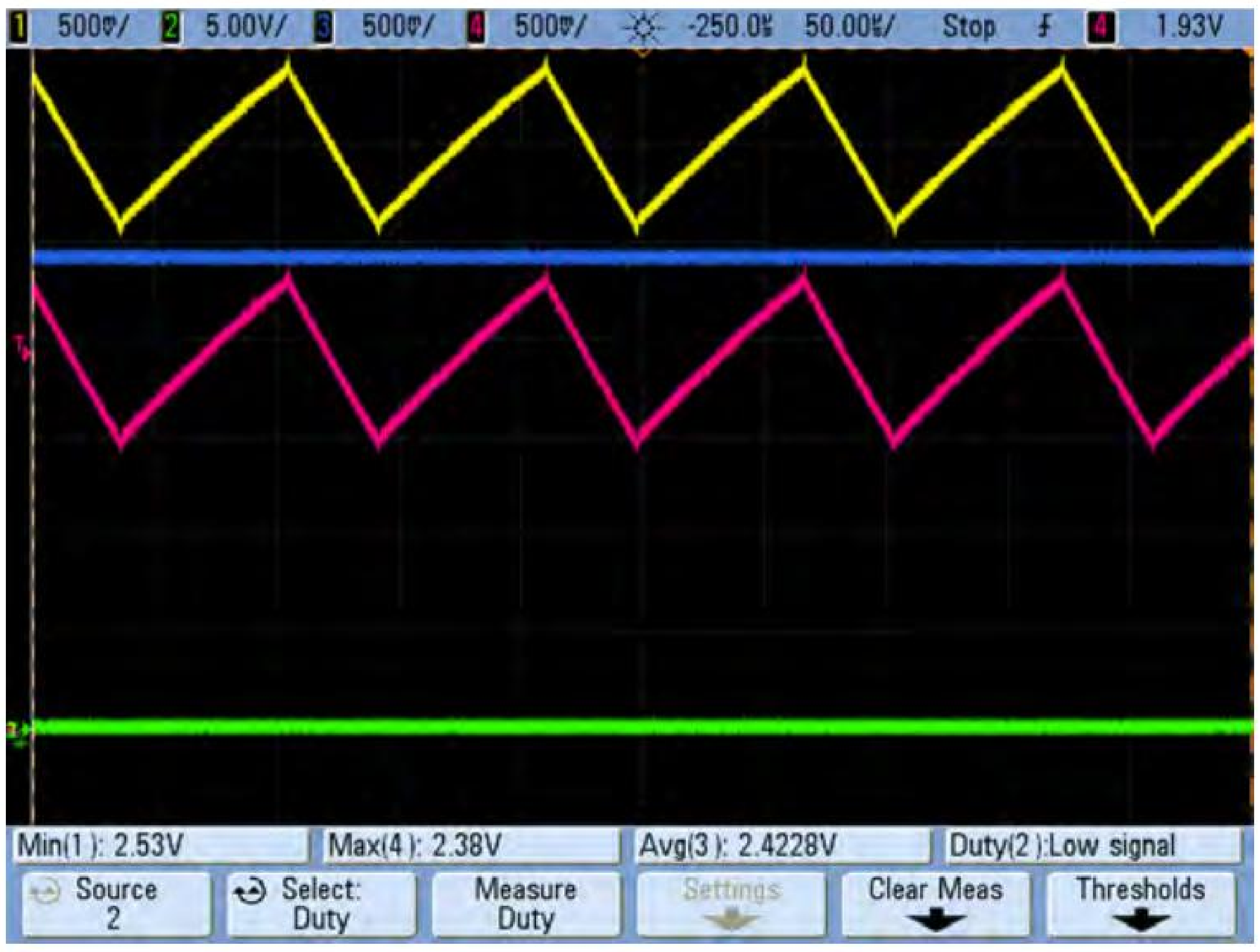


Fig.4 (right): the same voltages from pins 10 and 13 of IC1 are shown here but the blue trace now shows the reference voltage from speed pot VR2. Since it is below the yellow trace and above the mauve trace, no drive is applied to the tracks and the PWM output at pin 4 of IC2b, shown in green, is a flat line (ie, there is no PWM signal to tracks).

of Fig.4. The blue trace is the reference voltage from speed pot VR2. Since it is below the yellow trace and above the mauve trace, no drive is applied to the tracks and the PWM output at pin 4 of IC2b, shown in green, is a flat line.

Drive from both op amps (comparators) is fed to four of the six schmitt trigger inverter stages of IC2. IC2a and IC2f invert these signals and then drive LED1 and LED2, which have a common 220Ω current-limiting resistor. Hence, as the locomotive moves faster in the forward direction, LED1 lights up brighter (as it has a higher duty cycle) and similarly, the brightness of LED2 indicates the drive speed in the reverse direction.

The remaining four inverter stages are wired up in two series pairs, effectively forming buffers to square up the signals from IC1c and IC1d, and pass them to the inputs of integrated H-bridge IC3.

With IN1 and IN2 (pins 3 and 2) of IC3 both low, there is no output drive. With IN1 high, OUT1 (pin 6) is driven high while OUT2 (pin 8) is driven low. With IN2 high, OUT1 is driven low while OUT2 is driven high, reversing

Parts List

1 double-sided PCB available from the *EPE PCB Service*, coded 09103171, 143.5 × 50.5mm
1 flange mount ABS box, 125 × 80 × 35mm
1 panel label, 50 × 92mm
1 20kΩ single-turn horizontal PCB-mount trimpot (VR1)
2 No.4 × 5mm self-tapping screws
2 2-way 6.35mm PCB-mount terminal blocks (CON1,CON3)
1 PCB-mount DC socket, 2.1mm or 2.5mm ID (CON2)
1 6P4C RJ14 low-profile PCB-mount modular socket (CON4)
2 14-pin DIL sockets (optional)
10 PCB stakes (optional)

Semiconductors

1 TL074 quad JFET-input op amp (IC1)
1 MC14584 hex schmitt trigger inverter (IC2)
1 DRV8871 H-bridge IC (IC3)
1 78L05 100mA 5V linear regulator (REG1)
1 400V 4/6A vertical PCB-mount bridge rectifier (BR1)
2 3mm yellow LEDs (LED1,LED2)
1 3mm red LED (LED3)
2 3mm green LEDs (LED4,LED5)

Capacitors

3 1000µF 25V low-ESR electrolytic capacitors
1 10µF 6V tag tantalum capacitor
2 2.2µF 50V multi-layer ceramic capacitors
2 1µF 50V multi-layer ceramic capacitors
1 1µF 25V X7R SMD ceramic capacitor, 2012/0805 size
2 100nF 50V multi-layer ceramic capacitors
1 10nF 50V MKT capacitor

Resistors (all 0.25W, 1%)

2 10MΩ
2 47kΩ
1 22kΩ
1 18kΩ
1 10kΩ
1 3.3kΩ
3 2.2kΩ
1 1kW
3 220W

Additional parts for hand controller

1 PCB available from the *EPE PCB Service*, coded 09103172, 98 × 40.5mm
1 light grey ABS instrument case, 160 × 60 × 30mm
1 panel label, 51 × 94mm
1 6P4C RJ14 low-profile PCB-mount modular socket (CON5)
1 PCB-mount tactile switch with 22mm long actuator (S1)
1 100kΩ 16mm potentiometer with centre detent (VR2)
1 1MΩ 9mm vertical PCB-mount potentiometer (VR3)
1 button cap (for S1)
1 33mm black 1/4-inch shaft knob with white marker (for VR2)
1 11mm black 18 tooth spline plastic knob (for VR3)
4 No.4 × 5mm self-tapping screws
8 M3 nylon hex nuts
3 50mm lengths of light duty hookup wire
1 2m RJ14 to RJ14 telephone cable

Resistors (all 0.25W, 1%)

1 56kΩ
1 47kΩ
1 10kΩ

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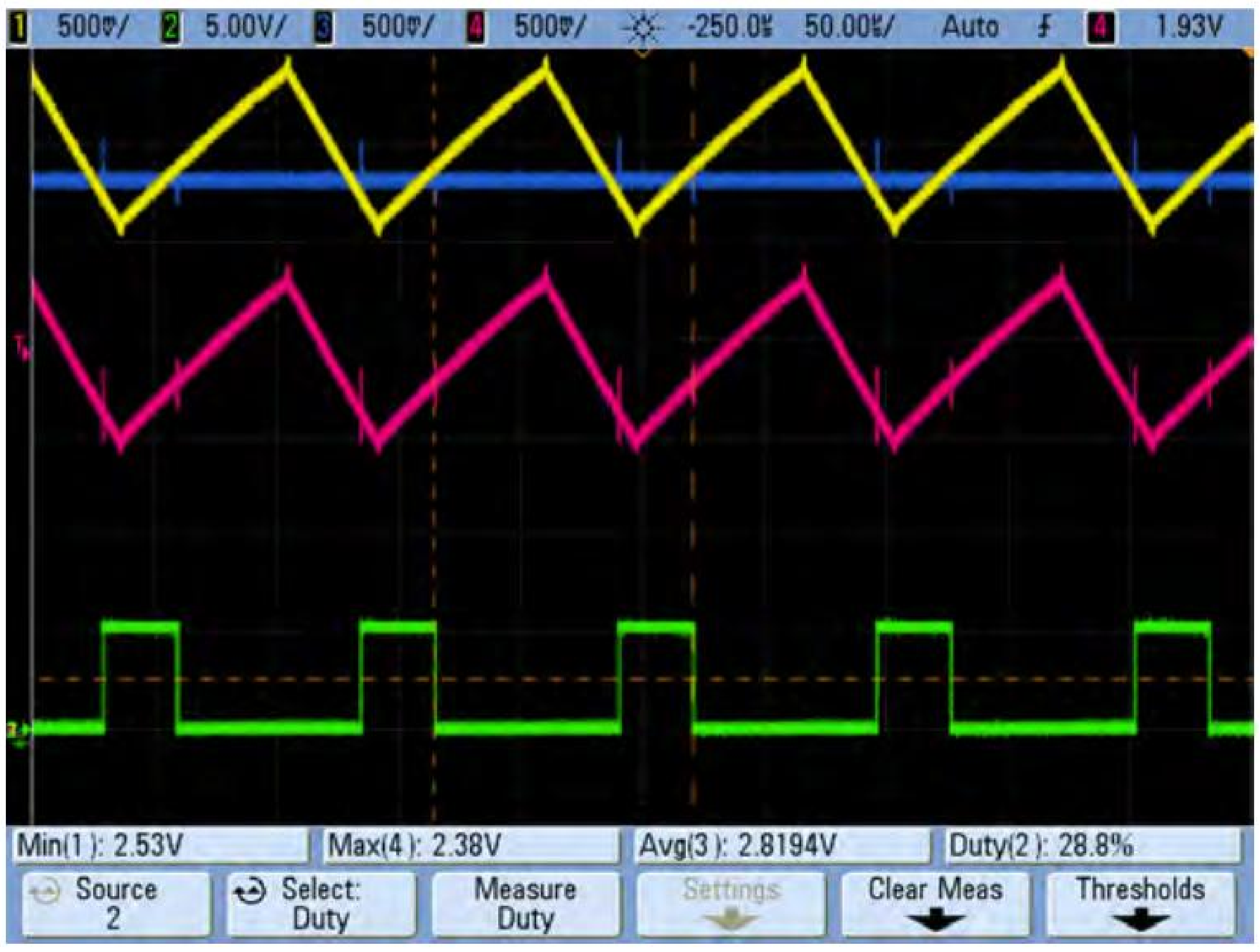
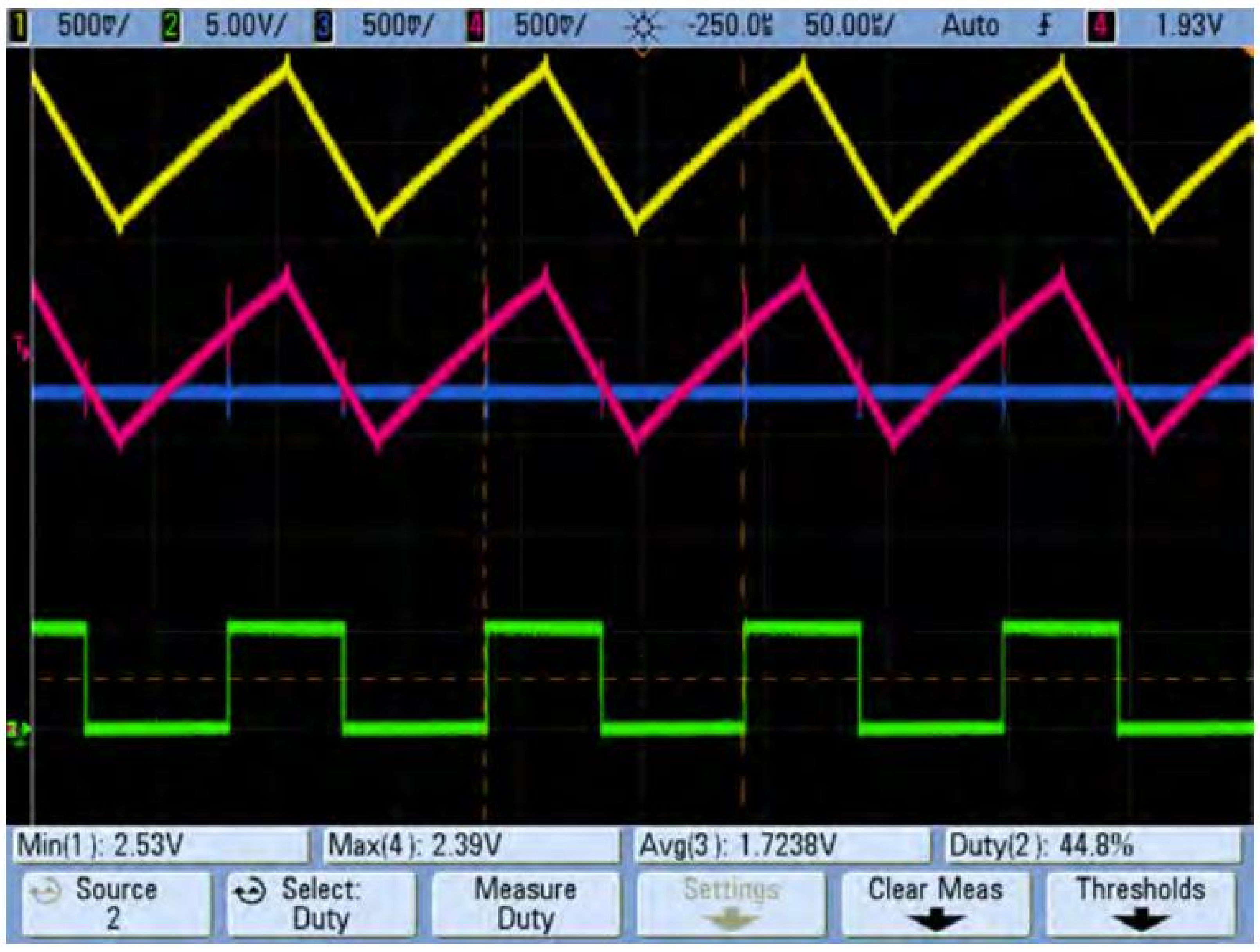


1 500mV/ 2 5.00V/ 3 500mV/ 4 500mV/ -250.0µs 50.00µs/ Auto 4 1.93V

Min(1): 2.53V Max(4): 2.38V Avg(3): 2.8194V Duty(2): 28.8%



1 500mV/ 2 5.00V/ 3 500mV/ 4 500mV/ -250.0µs 50.00µs/ Auto 4 1.93V

Min(1): 2.53V Max(4): 2.39V Avg(3): 1.7238V Duty(2): 44.8%

Fig.5 (left): the same traces as in Fig.4, but now the speed pot has been rotated clockwise, increasing the reference voltage (shown in blue). When the yellow waveform is below the blue reference voltage, the PWM output at pin 4 of IC2b, shown in green, increases to 5V and it drops back to 0V when the yellow and blue waveforms cross again. Thus, as the speed pot is rotated further clockwise, the PWM pulses at pin 2 of IC3 (IN2) increase in duty cycle.

Fig.6 (right): now speed pot VR2 has been rotated anti-clockwise past its centre position, so the reference voltage, shown in blue, has now dropped low enough to intersect with the mauve waveform. The green trace now shows output pin 6 of IC2c, which feeds input IN1 (pin 3) of IC3. Note that the positive edge of the PWM pulses is now delayed compared to the crossing point, due to the limited bandwidth of op amp IC1; however, the speed pot can still be used to adjust the PWM duty cycle.

the locomotive. And with IN1 and IN2 both high at the same time, both outputs are driven low to provide motor braking; however, that feature is not used in this circuit.

PWM output waveforms

We previously referred to the scope waveforms of Fig.3 and Fig.4, with the latter showing the condition where the speed control pot VR2 is centred, so there is no output at pin 2 of IC3 (IN2, green), nor at pin 3 (IN1, not shown).

In Fig.5, we have rotated VR2 partway clockwise and this has caused the control voltage (blue trace) to rise to 2.82V. As a result, pulses now appear at pin 2 of IC3 (IN2, green) with a duty cycle of 28.8%. You can see that the leading edges of these pulses correspond to the point where the yellow trace dips below the blue trace and the trailing edges are where they cross over again, so the higher the blue (control) voltage, the greater the applied duty cycle will be.

Fig.6 shows the situation with VR2 rotated anti-clockwise from its central detent, reducing the control voltage (blue trace) to 1.72V. The green trace now shows the voltage at pin 3 of IC3 (IN1) which has a duty cycle of 44.8% and the edges correspond to the points where the blue and mauve traces intersect.

H-bridge IC details

The internal block diagram of the DRV8871 IC is shown in Fig.7. It has four internal N-channel MOSFETs with parallel diodes that form the H-bridge which drives the motor; the circuit blocks to control the MOSFETs' gates; the charge pump to generate the required high-side and low-side gate drive voltages; and the various control and protection units within.

This IC has a current-limiting facility which both protects it from damage and also helps the unit withstand accidental short circuits across the track, as will inevitably happen on any model layout, particularly when a locomotive is derailed. The maximum output current depends on the value of R_{LIM} which connects between the I_{LIM} pin and ground. The IC is rated for up to 3.6A peak, so a current limit of around 3A, as set by R_{LIM} = 22kΩ is quite safe.

Should IC3 overheat due to extended high current delivery, it will automatically shut down until it has cooled sufficiently and then resume operation. IC3 also has an internal 'dead time' delay to prevent cross-conduction of its internal MOSFETs, which means that the driving circuitry can change the state of inputs IN1 and IN2 at any time without any chance of damaging the IC.

Referring back to Fig.2, IC3 also has an SMD ceramic 1µF bypass capacitor to help stabilise the output voltage and provide a relatively clean square wave for driving the motor. Note that IC3 has integral diodes between each output and the two supply rails, to

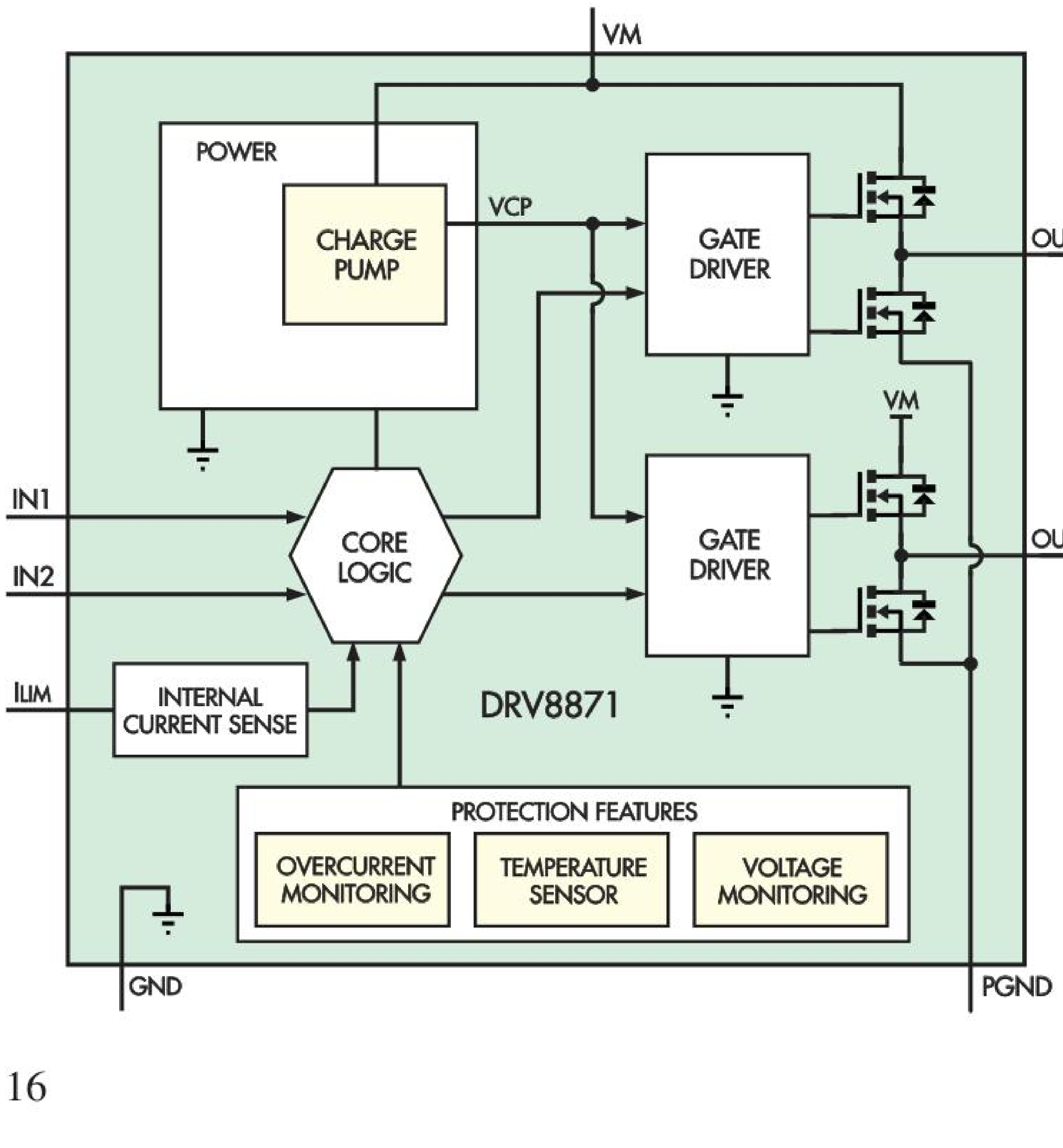


Fig.7: internal block diagram for the DRV8871 H-bridge IC. The internal MOSFETs are shown at upper-right; you can see the similarity in their connections to Fig.1. The IC also contains the boost circuitry to produce the required high and low-side drive signals to the MOSFET gates, control logic to prevent cross-conduction, plus current and temperature sensing and shutdown.

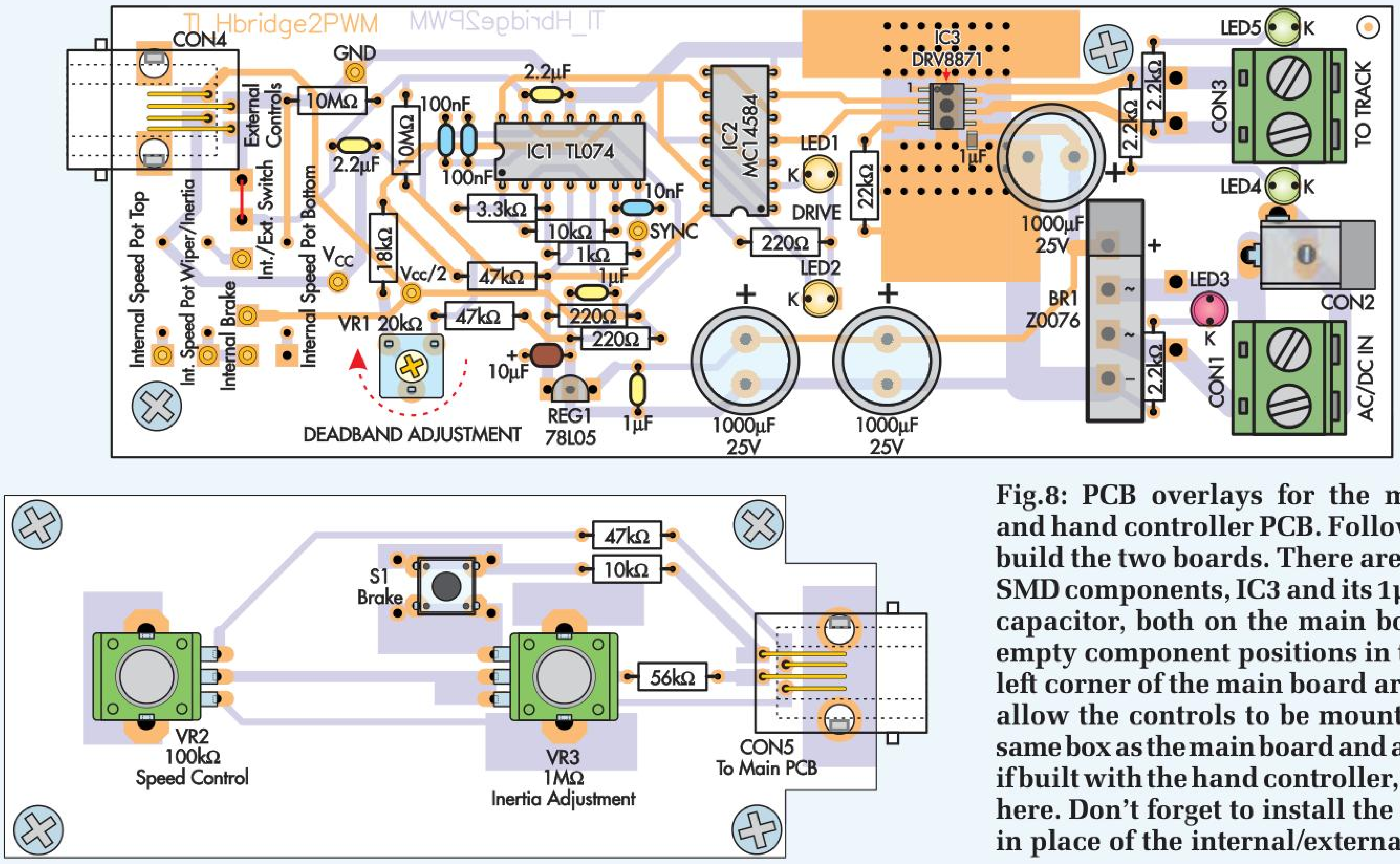

TI_Hbridge2PWM
CON4
GND
External Controls
10MΩ
2.2µF
100nF
10MΩ
2.2µF
100nF
IC1 TL074
IC2 MC14584
10nF
3.3kΩ
10kΩ
SYNC
1kΩ
18kΩ
VCC
Vcc/2
47kΩ
1µF
47kΩ
220Ω
220Ω
VR1 20kΩ
10µF
REG1 78L05
1µF
DEADBAND ADJUSTMENT
Internal Speed Pot Top
Int. Speed Pot Wiper/Inertia
Internal Brake
Int./Ext. Switch
Internal Speed Pot Bottom
LED1
K
DRIVE
22kΩ
220Ω
LED2
K
1000µF 25V
1000µF 25V
IC3 DRV8871
1µF
1000µF 25V
BR1 Z0076
LED5
K
2.2kΩ
2.2kΩ
CON3
TO TRACK
LED4
K
LED3
K
CON2
2.2kΩ
CON1
AC/DC IN
47kΩ
10kΩ
S1 Brake
56kΩ
VR2 100kΩ Speed Control
VR3 1MΩ Inertia Adjustment
CON5 To Main PCB

Fig.8: PCB overlays for the main PCB and hand controller PCB. Follow these to build the two boards. There are only two SMD components, IC3 and its 1µF bypass capacitor, both on the main board. The empty component positions in the lower left corner of the main board are there to allow the controls to be mounted in the same box as the main board and are left out if built with the hand controller, as shown here. Don't forget to install the wire link in place of the internal/external switch.

clamp any inductive spikes from the locomotive motor(s). It is purpose-designed for driving motors.

LED4 and LED5 are connected across the track outputs in opposite directions with 2.2kΩ current-limiting resistors and so normally echo the brightness of LED1 and LED2 respectively. However, if there is a short across the track, LED1/LED2 will still light, while LED4/LED5 will be off or dim. Note that LED4 and LED5 are located near the output terminal and are visible with the lid on the case.

Power supply

The power supply is quite simple and accepts either 10-15VAC or 12-19V DC. Actually, all the components should survive with a supply as high as 25V DC or 18VAC, should you wish to push it close to its limiting values.

LED3 is connected directly across the inputs and so will light solidly with a DC input or flicker with reduced brightness at 50Hz with an AC input. Either CON1, a 2-way terminal block, or CON2, a DC barrel connector can be used. We suggest you stick with the terminal block if your power supply is rated at more than 2A.

The input supply is rectified by bridge rectifier BR1 and this means that with a DC supply, the polarity of the connection is not important. The output of the rectifier is filtered with two parallel 1000µF capacitors, smoothing any ripples in the DC and also providing AC-to-DC conversion if required (in combination with BR1). The resulting DC is fed straight to the motor controller IC (IC3) and also to the input of 5V regulator REG1.

REG1 has a 1µF input bypass capacitor and 10µF tantalum output filter capacitor, and supplies IC2, IC3 and the two divider networks.

Construction

The *Stationmaster* is built on two PCBs. The main board is coded 09103171, measures 143.5 × 50.5mm and hosts most of the components. The hand controller board is coded 09103172, measures 98 × 40.5mm and is fitted with the components shown in the yellow box in Fig.2. Both of these boards are available from the *EPE PCB Service*.

Use the overlay diagrams in Fig.8 as a guide to construction, which is quite straightforward. The only slightly tricky component is IC3, which is only available in a surface-mount package, so start by soldering this. It has the additional twist that the underside of the IC features a metal pad which needs to be soldered to the PCB to provide sufficient heatsinking.

If you have a hot air rework station, all you need to do is apply a thin layer of solder paste to the central pad and eight pins for IC3, drop the IC in place (ensuring its pin 1 dot is oriented as shown in Fig.8) and then gently heat the IC until all the solder reflows. You can check that the solder underneath the IC has melted properly by examining it from the underside of the board through the three large vias positioned under IC3, once the board has cooled sufficiently.

If you don't have a hot air tool, we suggest you place a thin layer of solder paste (or at a pinch, flux paste) on the central pad for IC3, then position it as explained above and tack solder one of the eight pins using a regular soldering iron.

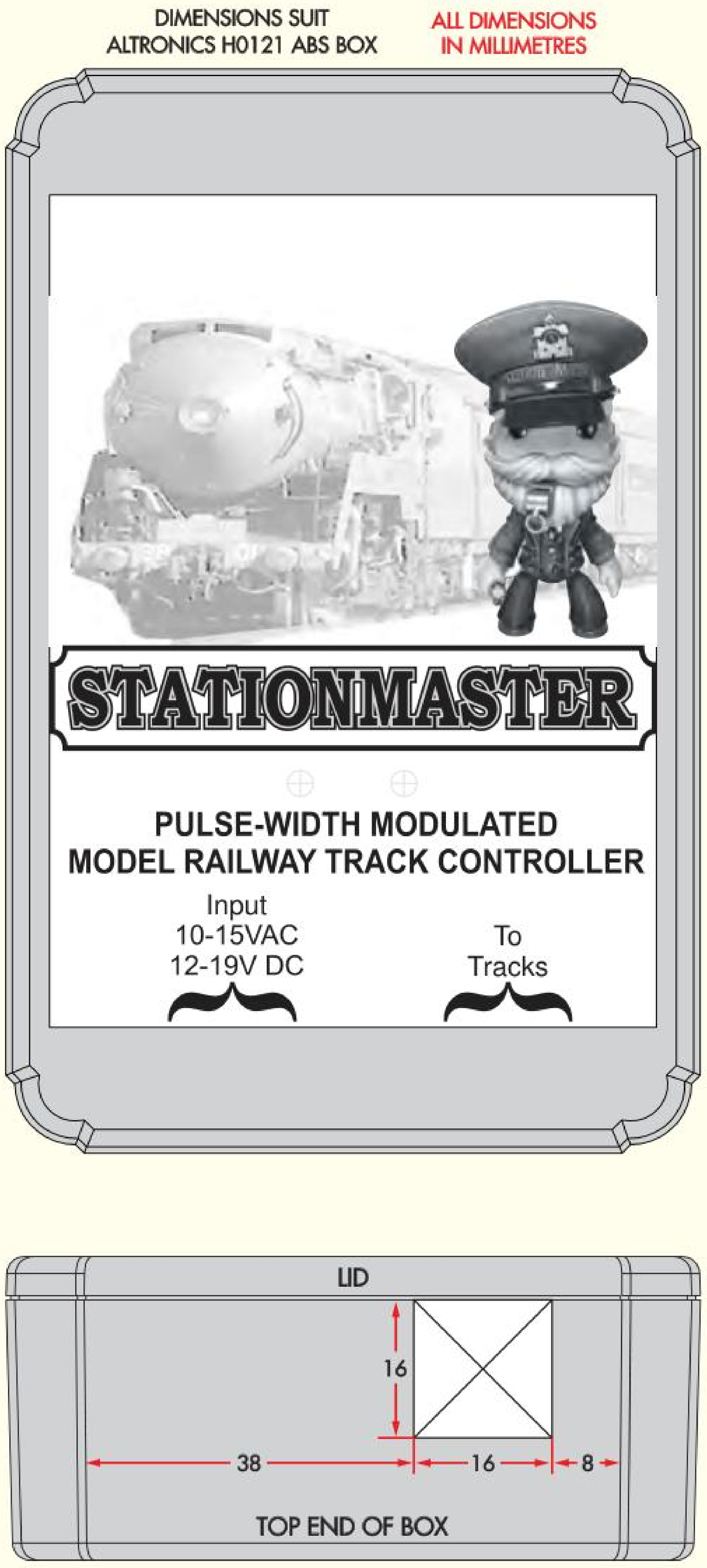
Check that the IC is sitting flat on the board and properly positioned over its pads, and then solder the remaining pins. Next, refresh the first pin which was tack-soldered. If any bridges form between its leads, clean them up using solder wick.

You can then flip the board over and melt some solder into the three large vias under the IC. Leave the iron in contact with this pad for a few seconds to ensure that the new solder remains molten and sufficient heat conducts through to the other side of the board to reflow the solder paste. That should do the trick and you can then remove any excess solder on the underside pad using a solder sucker or some solder wick.

There are also two small sets of SMD pads on either side of IC3, and the one to lower right is for the 1µF bypass capacitor. This is pretty easy to solder, simply tack solder one end, wait for the solder to cool, solder the other end (being careful to ensure the solder flows onto both the PCB pad and the end of the capacitor) and then apply fresh solder to the first joint.

Through-hole parts

With IC3 in place, the rest is pretty straightforward. Fit the 15 small resistors in the locations shown in Fig.8. It's a good idea to check the values with a


DIMENSIONS SUIT ALTRONICS H0121 ABS BOX
ALL DIMENSIONS IN MILLIMETRES
STATIONMASTER
PULSE-WIDTH MODULATED MODEL RAILWAY TRACK CONTROLLER
Input 10-15VAC 12-19V DC
To Tracks
LID
16
38
16
8
TOP END OF BOX

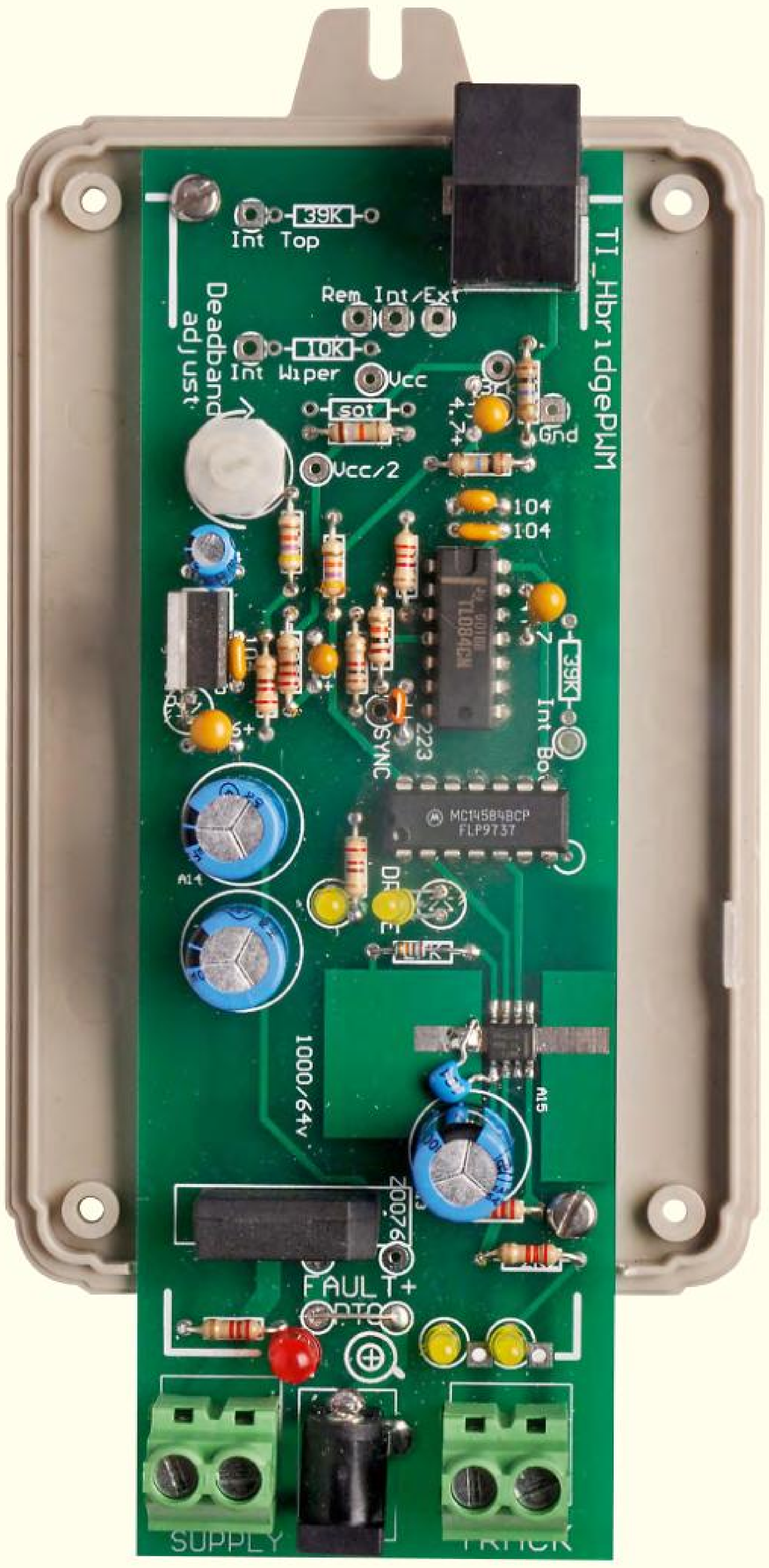


Fig.9: drilling and cutting diagrams for the main box. The top panel drilling template can also be used as the panel label.

DMM before fitting as the colour bands can be hard to identify accurately.

If you are using IC sockets, now is a good time to install them, making sure to orient the notches as shown in the overlay diagram. Otherwise, solder the other two ICs directly to the PCB, but be careful to make sure that you don't get them mixed up and that the pin 1 dot goes in the location shown.

Next, install all the small capacitors. The values are indicated on the overlay diagram. The capacitors of 1µF and above have a polarity (+) indicator, but do note that only the 10µF capacitor is actually polarised and this should have a matching + sign printed on its body, which must be lined up with that on the PCB.

LEDs 3-5 can now be fitted, taking care to orient them with the flat side of the lens/shorter lead (cathode) to the right/bottom of the board, where indicated with 'K' on the PCB overlay. These are pushed all the way down onto the PCB before being soldered and the leads trimmed.

You can now fit the PCB stakes if you want to; however, it isn't necessary and you can simply probe these pads with DMM leads if you want to troubleshoot the circuit.

Now mount trimpot VR1 and regulator REG1. You will need to crank REG1's leads to fit the solder pads, and make sure it goes in the right way around, with its flat face towards the nearest edge of the PCB. Note that a 7805 regulator can be used instead, and in this case, its metal tab faces the edge of the PCB.

Next on the list are DC connector CON2 and RJ12 connector CON4, both of which should be pushed all the way down onto the PCB before you solder their pins. You can then follow with terminal blocks CON1 and CON3, which must be fitted with their wire entry holes towards the right edge of the board.

Next, fit BR1, with its chamfered corner towards the top edge of the board. It should also have a + sign on the body of the device which you can line up with the polarity marker on the PCB. The three 1000µF capacitors can go in next, being careful to ensure that the longer (+) lead goes through the pad marked + in each case.

Now install LED1 and LED2. If you want these to be visible through the panel label on the lid of the box, fit them with the bottom of each lens 21mm above the top surface of the PCB.

However, these are really only necessary for diagnostic purposes, so you could just solder them flat on the PCB like the others. As before, the cathode side (shorter lead) is indicated in the overlay with a 'K', and this should line up with the flat side of the lens.

The main PCB is now complete and you can move on to building the hand controller.

Hand controller assembly

There aren't many components on this board. First, solder the three small resistors in place, then fit the RJ12 connector in the same manner as you did for the main board. Having done that, solder S1 and VR3 in place after making sure they have been pushed down fully onto the PCB.

For VR2, you can use a similar pot to VR3; however, it's better if you use the 16mm pot with centre detent, as specified in the parts list. In this case, the pot is be mounted on the case and attached to the PCB via three short (~50mm) flying leads. Refer to the photo above right to see how the wiring is done.

Completing the hand controller

The next step is to prepare the two cases to accept the boards. For the hand controller, this is simply a matter of drilling three holes in the lid for the two pots and pushbutton shaft to poke through.

You can download the panel label artwork from the *EPE* website and use this as a drilling template; or copy Fig.10. The hole for the 9mm pot should be drilled to 7mm, and 8mm for the 16mm potentiometer. Ideally, you should also drill a 3mm hole for the latter pot's locking tab, although you can simply snap this off (but then you will need to do its nut up tight to stop it rotating).

Having done that, print and affix the panel label (see the link below Fig.10 for suggestions on how to do this) and cut out the holes with a sharp hobby knife; there's no need to make a hole for the pot's locking tab, as this will not protrude through the case.

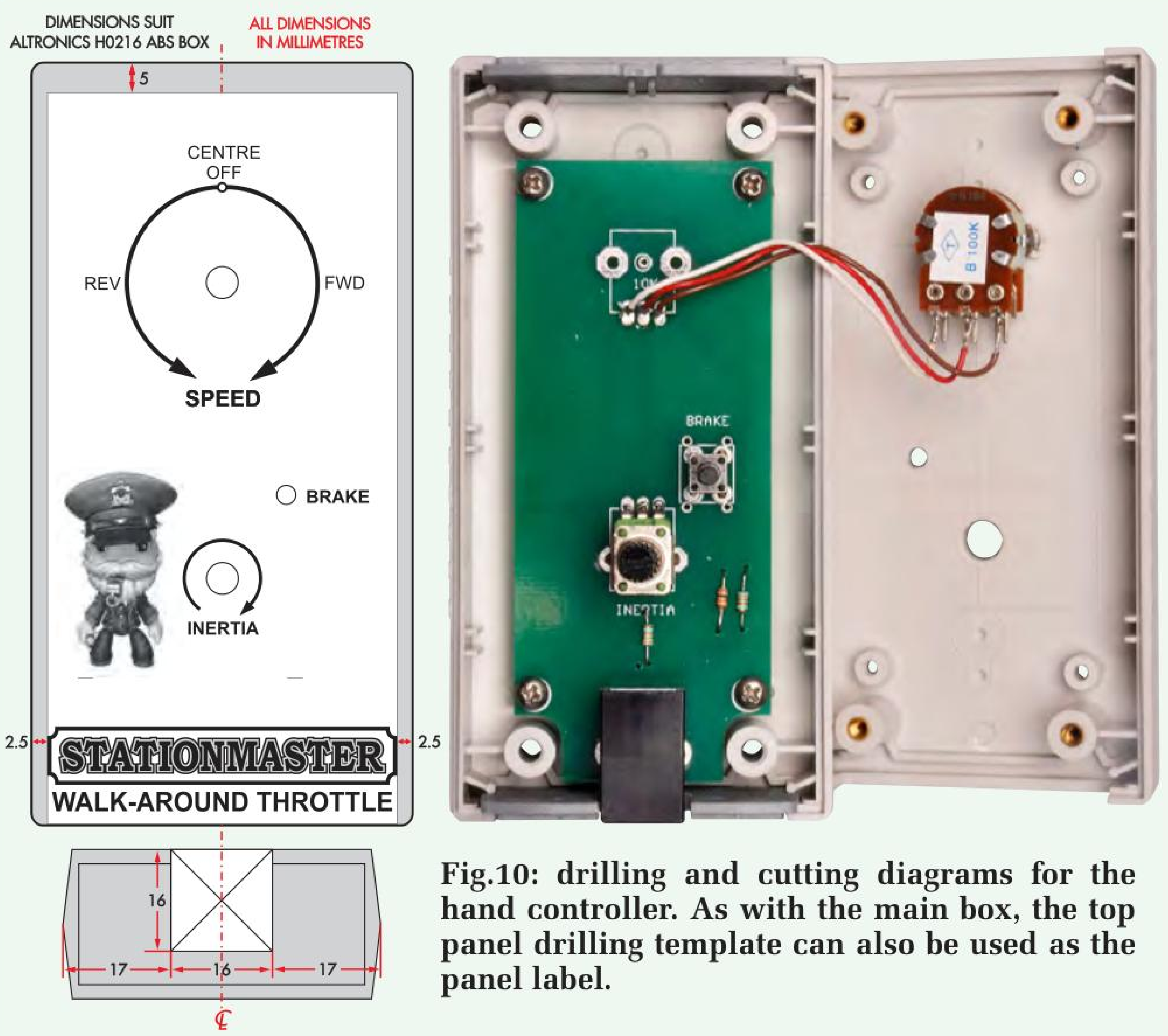


Fig.10: drilling and cutting diagrams for the hand controller. As with the main box, the top panel drilling template can also be used as the panel label.

If you want to make your own label for either of the cases we have a short description on our website on printing A4-sized synthetic sticky labels here: www.siliconchip.com.au/Help/FrontPanels

Now cut and/or file a rectangular hole in the case end panel, as shown in Fig.10. You can then insert this into the appropriate slots and affix the hand controller PCB to the integral posts in the bottom of the case using four small self-tapping screws. Note though that you need to place two M3 nylon nuts on top of each of these posts before inserting the screws; these act as spacers to get the modular socket to just the right height.

It's then simply a matter of inserting the other end panel into the case, placing the lid on top, using the four supplied screws to join the two halves of the case together and then attaching the two knobs and the button cap for S1. The knob for VR3 and the button cap for S1 are simply pressed on and held by friction (note that you will need to use the grub screw to attach the knob for VR2).

Completing the main unit

Now to complete the main unit. First, you need to cut or file down the rim around the lid of the case so that when you attach the PCB later, the part which projects out the side will not be fouled by this rim. See the photo adjacent to Fig.9 for details.

Having done that, the next step is to make the cut-out for the modular socket in the side of the case. Fig.9 shows the detail. The only remaining holes that need to be made are for LED1 and LED2, assuming you've decided to install them with long leads so that they can be seen with the lid on. The positions for these 3mm holes are shown in Fig.9.

Now affix the panel label, using the same technique as for the hand controller, making sure the 'Motor Drive Present' text goes just below the two holes if you have drilled them. The label should be oriented so that the logo is near the cut-out for the modular socket.

Then attach the PCB to the lid using two short self-tapping screws and check that the two halves of the case fit together properly and the top of the LED lenses poke through the hole (if you've made them).

But before you actually put the case together, we need to do some testing and adjustment.

Test and set up

Plug the hand controller into the main board using a 4-wire telephone cable and centre the speed pot while the inertia pot should be fully anti-clockwise. Adjust trimpot VR1 on the main board to be fully clockwise.

Apply power to the main board via CON1 or CON2 and check that LED3 lights. The other LEDs should be off. If any of the other LEDs light up, switch off and check for faults. Using IC3's ground plane as the 0V reference, check for 4.5-5.5V at the V_{CC} test point and half that at the $V_{CC}/2$ test point. If you have a frequency meter, measure the frequency at the SYNC test point. It should be in the range of 8-10kHz.

Measure the AC voltage across the terminals of CON3. You should get 0V. Now slowly rotate VR1 anti-clockwise until LED1 and/or LED2 light up, then back off slightly until both LED1 and LED2 are off. Check again that you have 0V at CON3.

You can now slowly rotate speed pot VR2 in one direction. If rotating clockwise, LED1 and LED4 should both light up and get brighter as you turn the pot further. If rotating anticlockwise, LED2 and LED5 should both light up and get brighter as you turn the pot further.

Now rotate the inertia pot clockwise and the above should still hold true, but you should notice that the rate of change of LED brightness has been reduced. With the speed pot fully at one stop, hold down brake switch S1 and check that LED1, LED2, LED4 and LED5 all switch off in fairly short order and return to their previous states once you release it.

As a final test, you can hook up the CON3 terminals to a pair of train tracks and check you can control the speed and direction of a locomotive on those tracks as expected. If it moves in the opposite direction to what you intend, simply swap the connections at CON3.

Final assembly and usage

Now that you've confirmed it's working, you can join the two halves of the box with the supplied screws and integrate the controller into your layout.

Note that pressing and holding the brake button will bring everything to a halt very quickly; practice will allow you to tap S1 to slow a locomotive, which will return to set speed when you release it.

If you do need to use S1 for emergency braking, remember to set speed potentiometer VR2 to its central position (easy if you've used a pot with centre detent) before releasing S1 in order to prevent the locomotive from moving again when S1 is released.

RJ12 adaptors can be purchased and placed along a loom cabled around the layout so that the hand control can be unplugged and moved to a different location as you operate.

The speed set at the time of unplugging will be maintained for a period and will slowly diminish over time until control is re-established, which might cause a rapid return to the former speed. It's best to set the inertia control fairly high before plugging the controller back in to avoid this.

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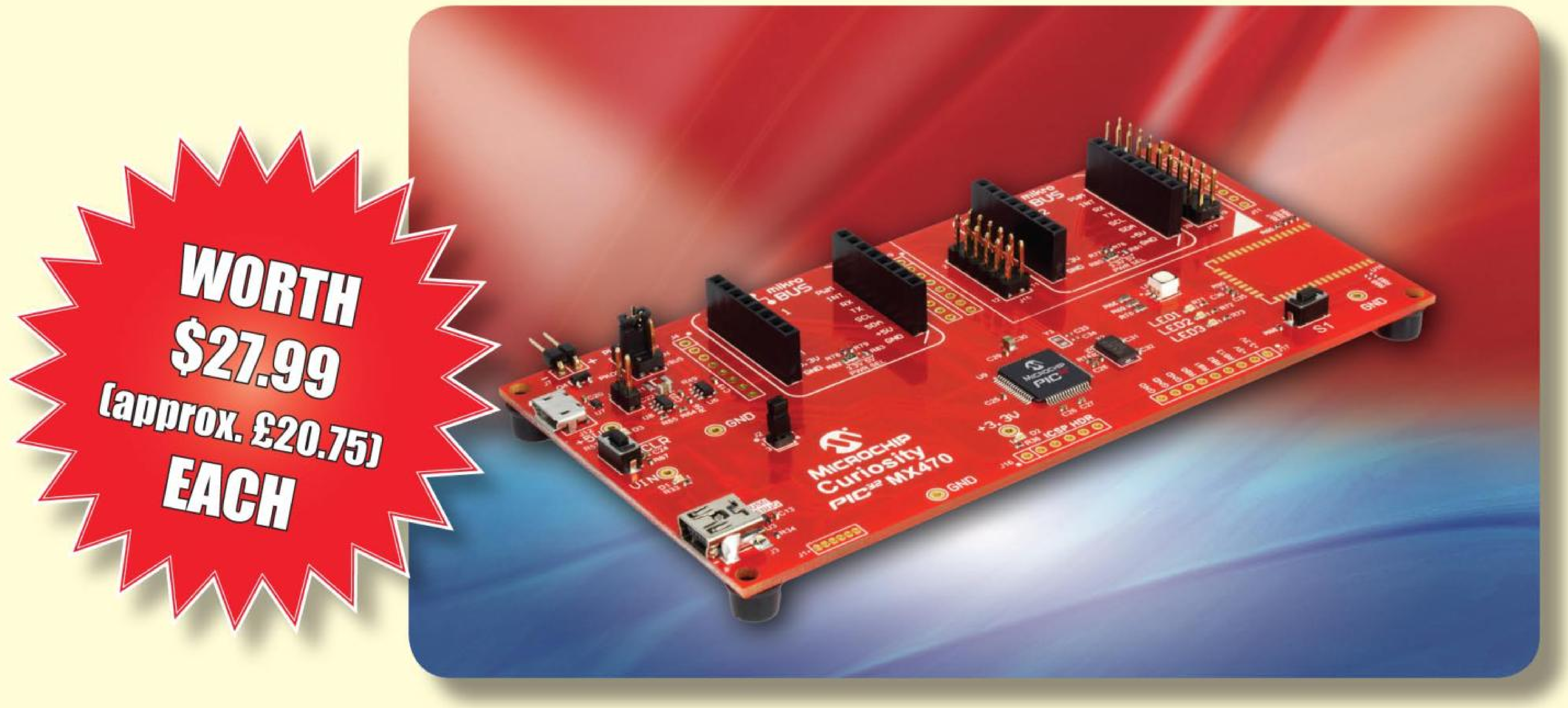

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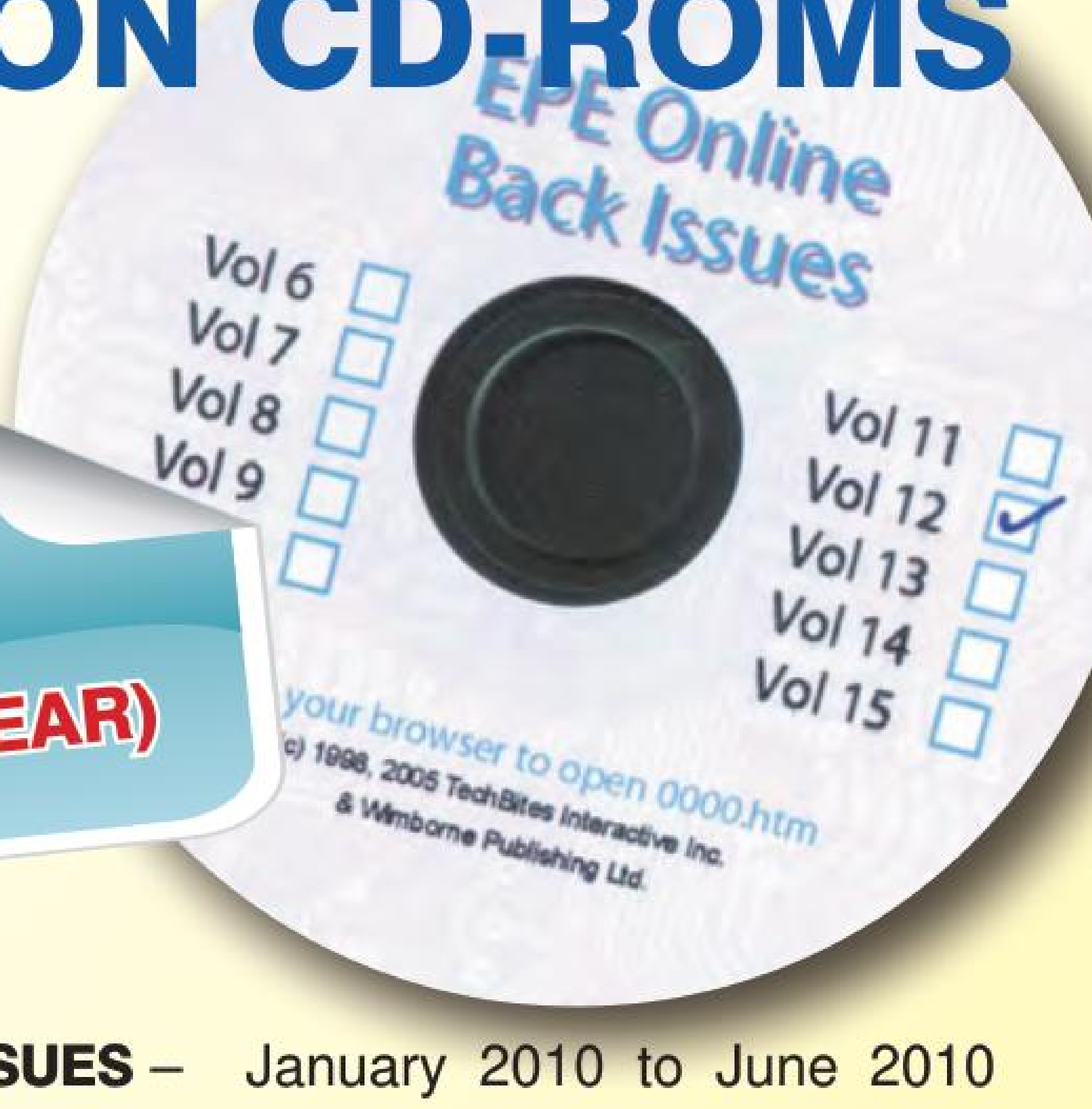
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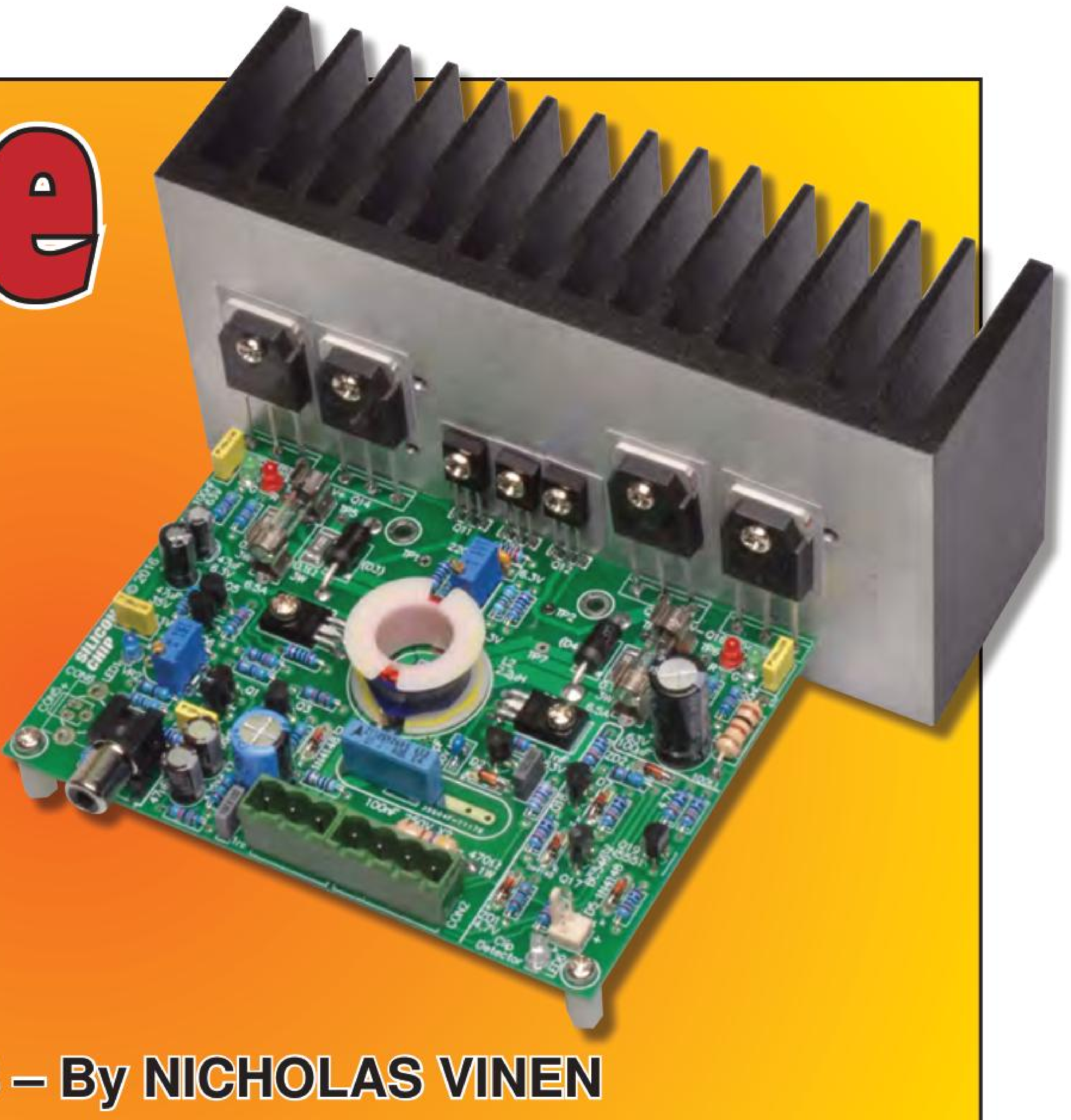


Part 3 – By NICHOLAS VINEN

In this third instalment, we provide the SC200's performance details. We also describe the required power supply, the testing and set-up procedure and how to build lower-power versions of the amplifier.

The SC200 is our new workhorse audio amplifier module, and while it doesn't have the extremely high performance of the very best designs, it's still more than comparable with most brand-name Hi-Fi amplifiers and it also has power aplenty. It's also easier to build and the parts cost significantly less than many SMD designs.

Fig.7 shows where the SC200 has the biggest advantage over a 14-year old design (the SC480) and that's in power output. The first thing you may notice is that below 10W, the total harmonic distortion of the SC200 is slightly higher than the SC480 but that's simply because it has more gain.

Since both designs use BC557 transistors at the input, their absolute noise figures are very similar, but since the SC200 delivers a lot more power, it needs more gain and this also amplifies the noise more.

Hence, while the SC200's signal-to-noise ratio relative to full power is 1dB better than the SC480, the noise at a particular power level will be slightly higher.

Having said that, at power levels above 10W the SC200 delivers significantly lower distortion. The SC480 runs into clipping at around 55W for 8Ω loads and 70W for 4Ω loads, while the SC200 delivers a clean output up to power levels of 135W for 8Ω loads and 200W for 4Ω loads.

Music power (ie, for short bursts such as percussion instruments) is even higher, at around 150W into 8Ω and 250W into 4Ω. So the SC200 has substantially more power output than the olde SC480.

Fig.8 shows distortion for the new SC200 and old SC480 designs at the same power level, into the same resistive loads and over the entire audible frequency band. We've used the plots for the TO-218 (plastic package transistor) version of the SC480 to be fair, since it is the more modern of the two designs that were originally presented, and it gave slightly better performance.

As you can see, the shapes of the distortion curves for both designs are very similar, but at the power levels used here, the SC480 has about 1/3 the distortion at all frequencies.

Note that we have filtered out some of the noise with a 30kHz bandwidth, to allow us to better see the harmonic distortion; the SC480 article doesn't state what bandwidth was used so it's difficult to make an 'apples-to-apples' comparison.

We have shown the projected high-frequency distortion with dotted lines, taking into account the fact that the limited bandwidth will filter out some of the higher harmonics for those frequencies.

Given that noise has less of an effect on the distortion measurements at higher frequencies, because it becomes a less significant proportion of the rising THD+N, this does suggest that the SC200 will have noticeably lower distortion at higher frequencies, at least into 8Ω loads, and should sound slightly better when driving 4Ω loads too.

Fig.9 compares the frequency response of both amplifiers at 10W into an 8Ω load. The frequency response of the SC480 is –1.8dB at 20Hz and –1.6dB at 20kHz. By comparison, the SC200's response is astonishingly flat at just –0.06dB at 10Hz and –0.13dB at 100kHz.

That more extended bass response will certainly be apparent if your CD player and your discs have very low bass signals (such as those from a pipe organ with 64-foot pipes!) and if your loudspeakers have the bass performance to match. At the other end of the spectrum, you will need young ears able to hear up around 20kHz and good speakers and program source to be able to notice the difference.

Power supply

In the power supply for the SC200 we rectify the output of a 40-0-40V toroidal transformer and feed it to a 6 × 4700µF capacitor bank to generate the nominal ±57V supply rails. The power supply PCB also carries optional circuitry to derive a ±15V preamplifier supply from a second 15-0-15 transformer, or a secondary winding on the main transformer.

The full circuit for the power supply is shown in Fig.10. This shows component values for the full-power rated supply, but also for a lower-voltage version which will reduce the power output slightly, to 75W into 8Ω loads and 110W into 4Ω loads.

There isn't a great deal to the power supply circuit. An external 35A bridge rectifier converts the AC from the transformer into pulsating DC, which is used to charge the two large capacitor banks. LED1 and LED2 act as bleeders, to discharge this bank after switch-off and also show when the supply is live.

A separate 1A on-board rectifier comprising diodes D1-D4 and two 2200µF capacitors converts the 15-0-15V AC output of the secondary windings to around ±20V DC, which is then fed to a pair of linear regulators to produce the ±15V rails for the preamplifier (or whatever other circuitry you need to power within the chassis).

The power supply PCB overlay is shown in Fig.11. The preamplifier regulator section at right can be cut off if you don't need it, or want to mount it elsewhere. The output of the bridge rectifier is connected via three spade quick-connect terminals, while two sets of DC outputs are provided on either side, making it easier to build a stereo amplifier.

Although we show a couple of wire links on this PCB, production boards should have WIDE top layer tracks joining those points, so fitting these wire links is not necessary. Check your board to verify this before starting assembly. The parts list for building the power supply is included later on in this article.

Lower-power amplifier module

If you want to build the lower-voltage power supply, using a 30-0-30VAC transformer which gives around ±42V DC, you need to make some slight changes to the amplifier modules.

The most important change is that the 22kΩ resistor between the collector of Q7 and ground (to its right on the PCB) must be changed to 15kΩ.

It's also a good idea to change the two 6.8kΩ resistors at the collector of Q6 (one to its left and one below VR2) to 4.7kΩ; however this is less critical and it will probably work OK with the original values.

Building the power supply

You'll need to build a power supply before you can test the amplifier module(s). Use the overlay diagram in Fig.11 as a guide to fit the components to the PCB, which is available from the *EPE PCB Service*, coded 01109111. Note that the power supply module kit is available from Altronics; Cat K-5168 (does not include transformer – you choose which one you want).

Assuming you do want the low voltage outputs, fit the four 1N4004 diodes (D1-D4), orienting them as shown. Then install the two 3-terminal regulators. You will need to bend their leads down by 90° so that they fit the PCB pads with the tab mounting hole lined up correctly. Attach each regulator to the board using an M3 x 6mm machine screws, shakeproof washer and nut, taking care not to get the two different types mixed up. Solder the leads *after* the screws have been tightened.

The two LEDs can go in next. These sit flush against the PCB with the flat side of the lenses oriented as shown on the overlay.

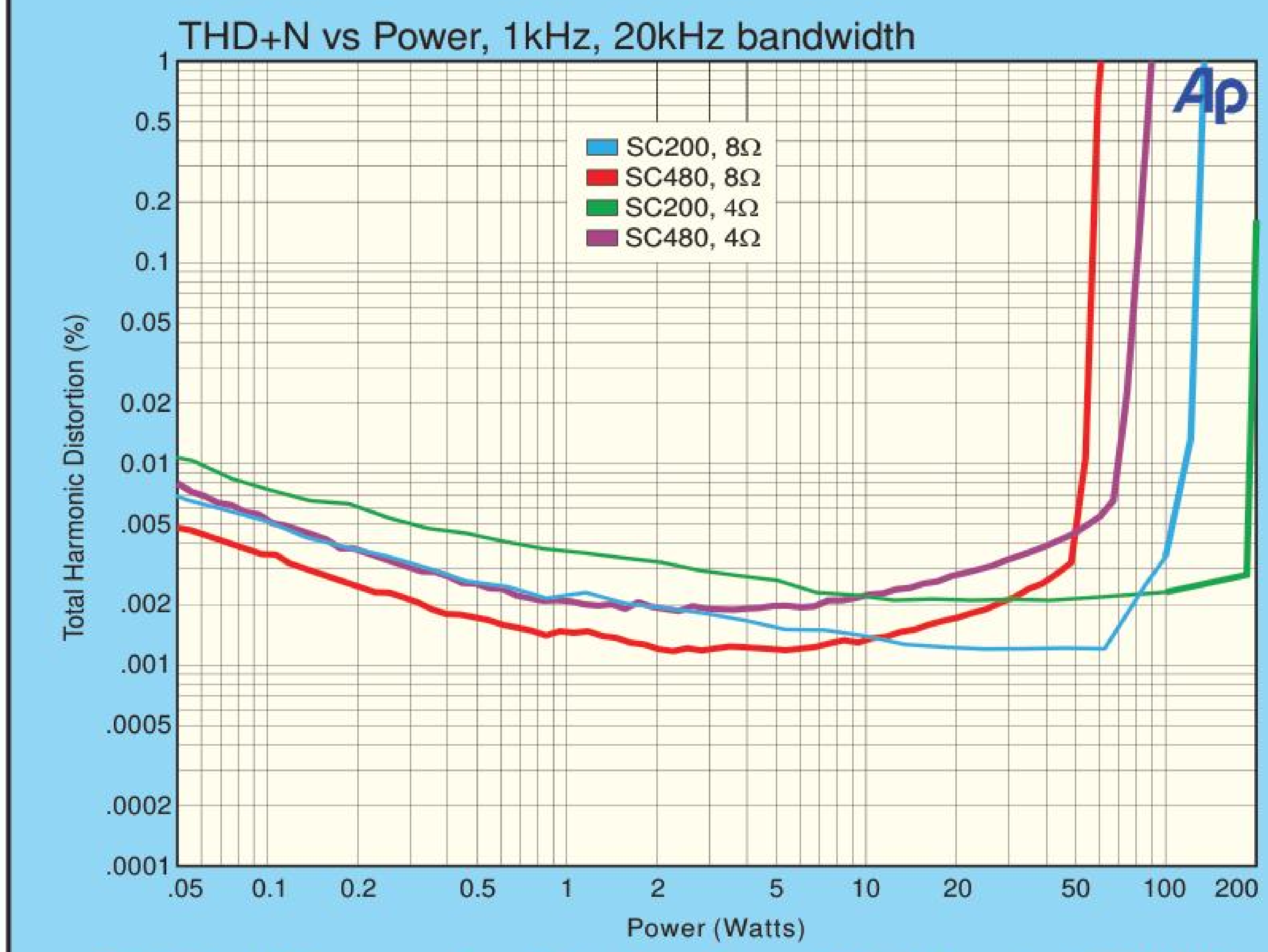


Fig.7: total harmonic distortion from 50mW up to 200W for the new SC200 amplifier, compared to the older SC480 design. Distortion is slightly higher below 10W due to the increased gain and thus noise, but significantly improved for powers above 10W and maximum power is much higher.

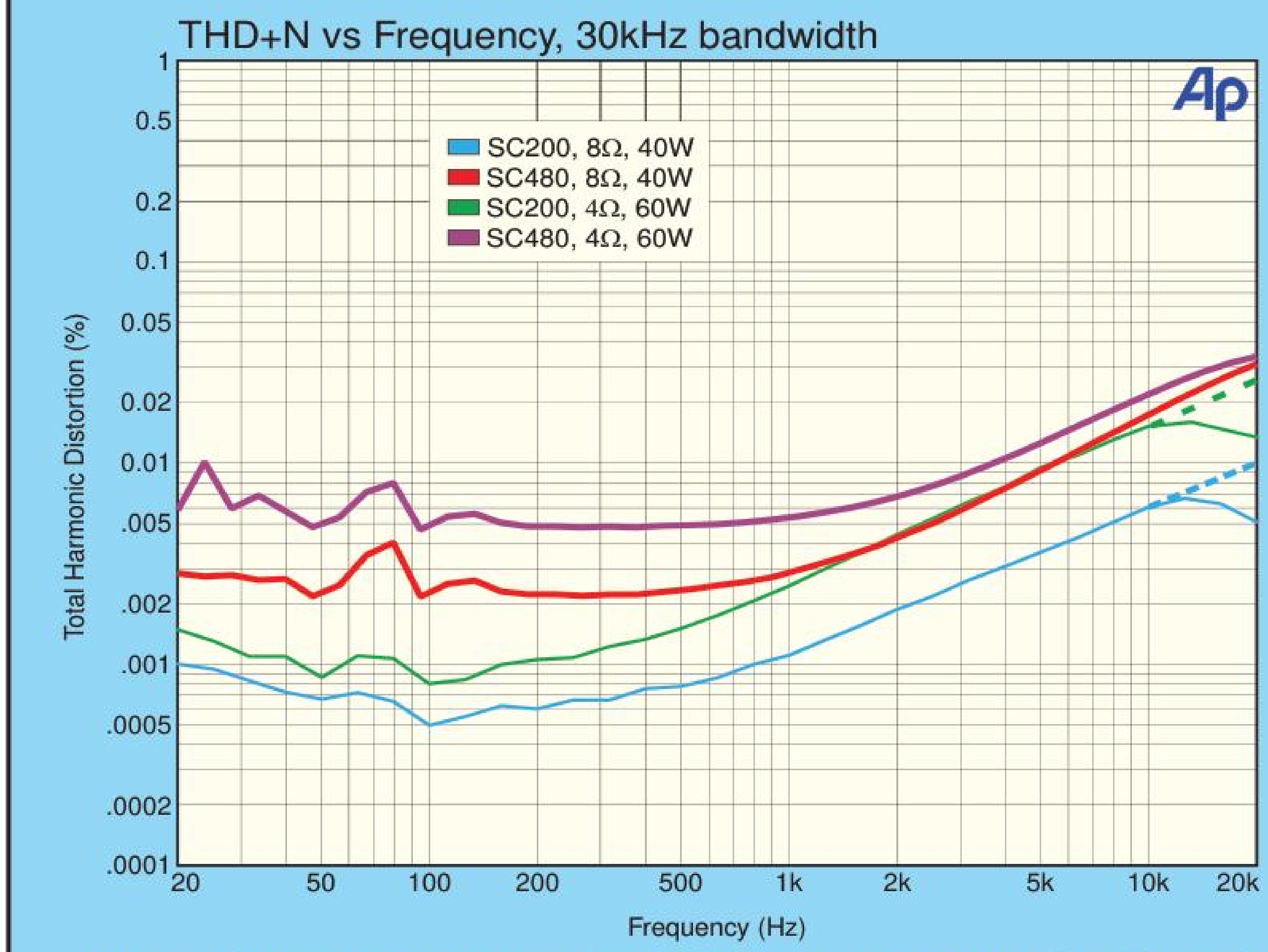


Fig.8: distortion versus frequency at 40W (8Ω load) and 60W (4Ω load). These power levels are the nominal output powers for the SC480 and this allows a direct comparison. As you can see, the distortion of the SC200 is lower, especially for 8Ω loads.

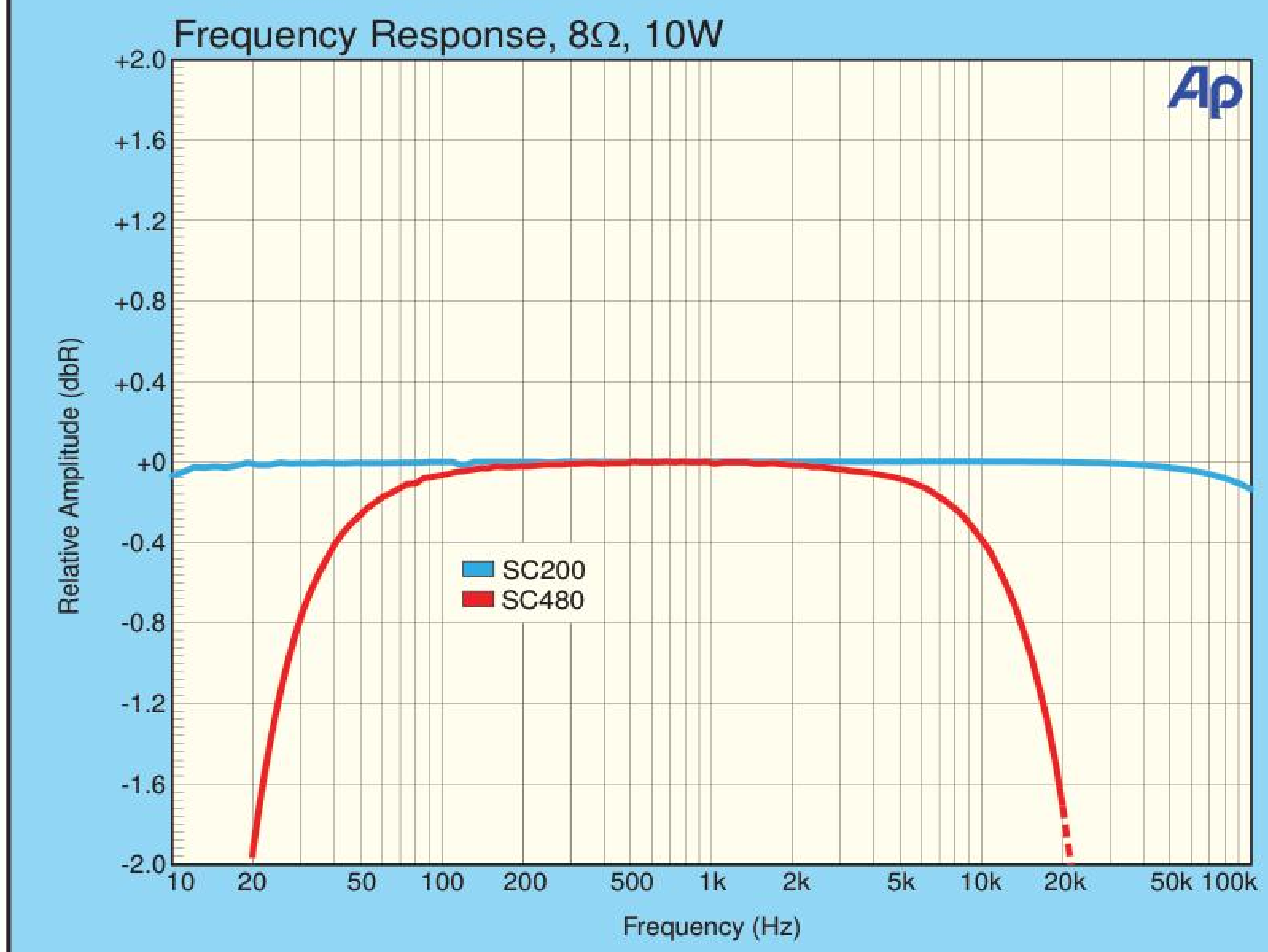


Fig.9: the frequency response of the SC200 is almost ruler-flat over the range of 10Hz-100kHz and should result in greatly extended bass, compared to the SC480.

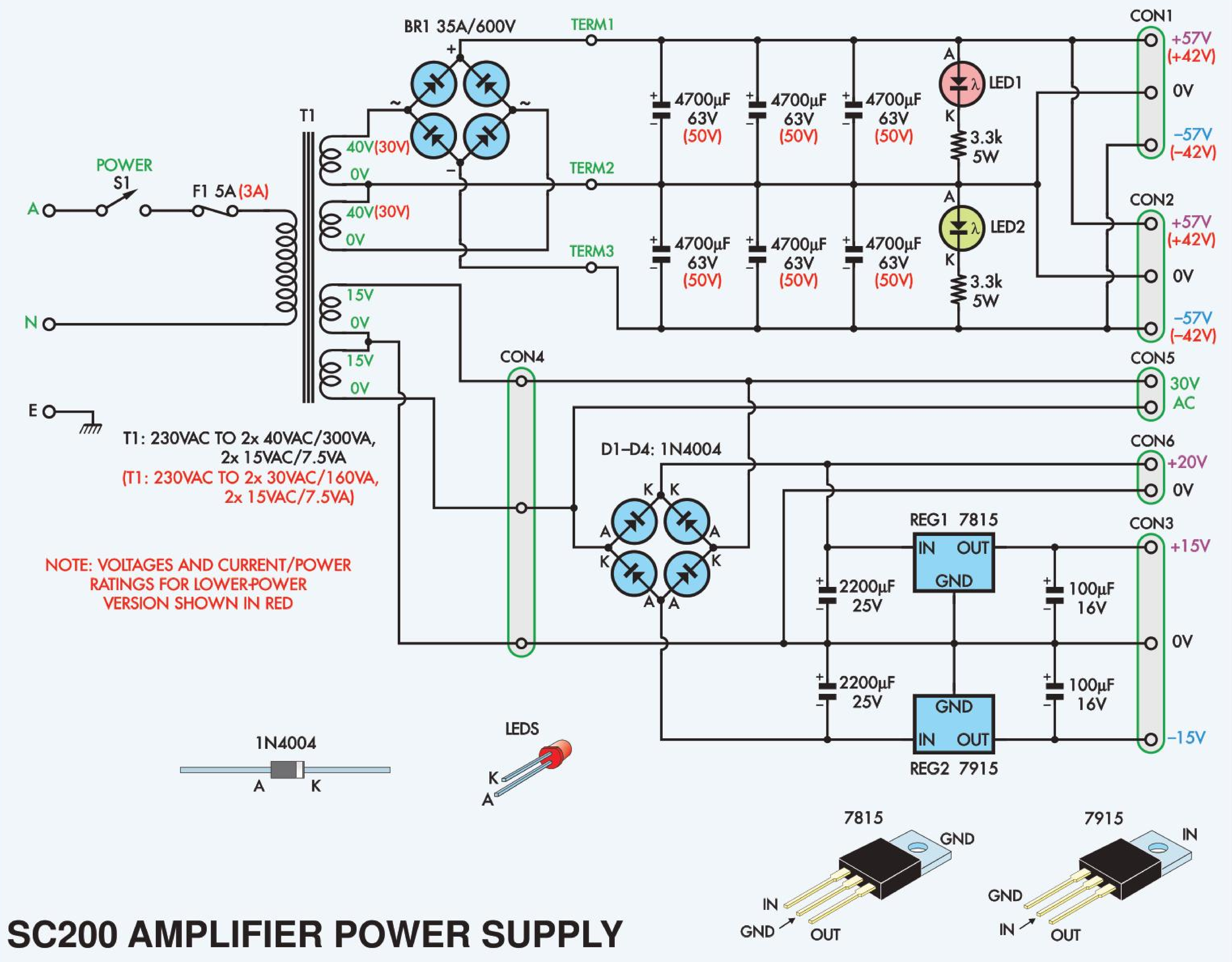

BR1 35A/600V
TERM1
TERM2
TERM3
T1
POWER
S1
F1 5A (3A)
A
N
E
40V (30V)
0V
40V (30V)
0V
15V
0V
15V
0V
4700µF 63V (50V)
4700µF 63V (50V)
4700µF 63V (50V)
4700µF 63V (50V)
4700µF 63V (50V)
4700µF 63V (50V)
LED1
LED2
3.3k 5W
3.3k 5W
CON1
+57V (+42V)
0V
–57V (–42V)
CON2
+57V (+42V)
0V
–57V (–42V)
CON4
CON5
30V AC
CON6
+20V
0V
CON3
+15V
0V
–15V
D1–D4: 1N4004
REG1 7815
IN OUT GND
REG2 7915
GND IN OUT
2200µF 25V
2200µF 25V
100µF 16V
100µF 16V
T1: 230VAC TO 2x 40VAC/300VA, 2x 15VAC/7.5VA
(T1: 230VAC TO 2x 30VAC/160VA, 2x 15VAC/7.5VA)
NOTE: VOLTAGES AND CURRENT/POWER RATINGS FOR LOWER-POWER VERSION SHOWN IN RED
1N4004
A K
LEDS
K
A
7815
GND
IN
GND
OUT
7915
IN
GND
IN
OUT
SC200 AMPLIFIER POWER SUPPLY

Fig.10: complete power supply circuit for the SC200. Depending on which transformer is used, the main DC rails are either ±57V, giving 135W into 8Ω and 200W into 4Ω, or ±42V, giving 75W into 8Ω and 115W into 4Ω.

Follow these with the two 3.3kΩ 5W resistors. These should be stood off the board by about 2mm, to allow the air to circulate beneath them for cooling (use a cardboard spacer during soldering).

The two 5-way screw-terminal connectors are made by dovetailing 2-way and 3-way blocks together. Be sure to fit these assemblies with the wire entry holes facing towards the adjacent edge of the PCB.

The two 3-way terminal blocks for the ±57V (or ±42V) outputs can then go in. Alternatively, instead of fitting these blocks, you can solder the DC supply leads directly to the PCB pads if it will be mounted right next to the amplifier modules.

The three quick-connect (spade) terminals are next on the list. If you are using PCB-mount connectors, simply push the pins through and solder them in place. It will take a while to heat the connectors so that the solder will 'take'. However, be careful not to overdo it, as the solder could 'wick' through the hole and onto the spade section.

If you are using 45° chassis spade lugs instead, screw them down tightly using M4 machine screws, nuts and washers – see Fig.12. If you can't get single-ended chassis lugs, cut one side off double-sided lugs.

Finally, fit the electrolytic capacitors, starting with the two 220µF units and finishing with the six large 4700µF units. Be sure to orient them correctly and make sure that they all sit flush with the PCB.

If building the lower power version, you'll probably need to crank out the capacitor leads to suit the board and it would also be a good idea to apply a little neutral-cure silicone sealant around the base of the capacitors so they aren't supported by the leads alone.

Cabling

Note that it's important to use the thickest wire you can easily fit into the terminal blocks and to keep the wiring as short and as tight as possible.

Each set of three wires from the power supply to the amplifier module should be tightly coupled by twisting them together and/or covering the bundle with a length of heatshrink tubing – ideally both.

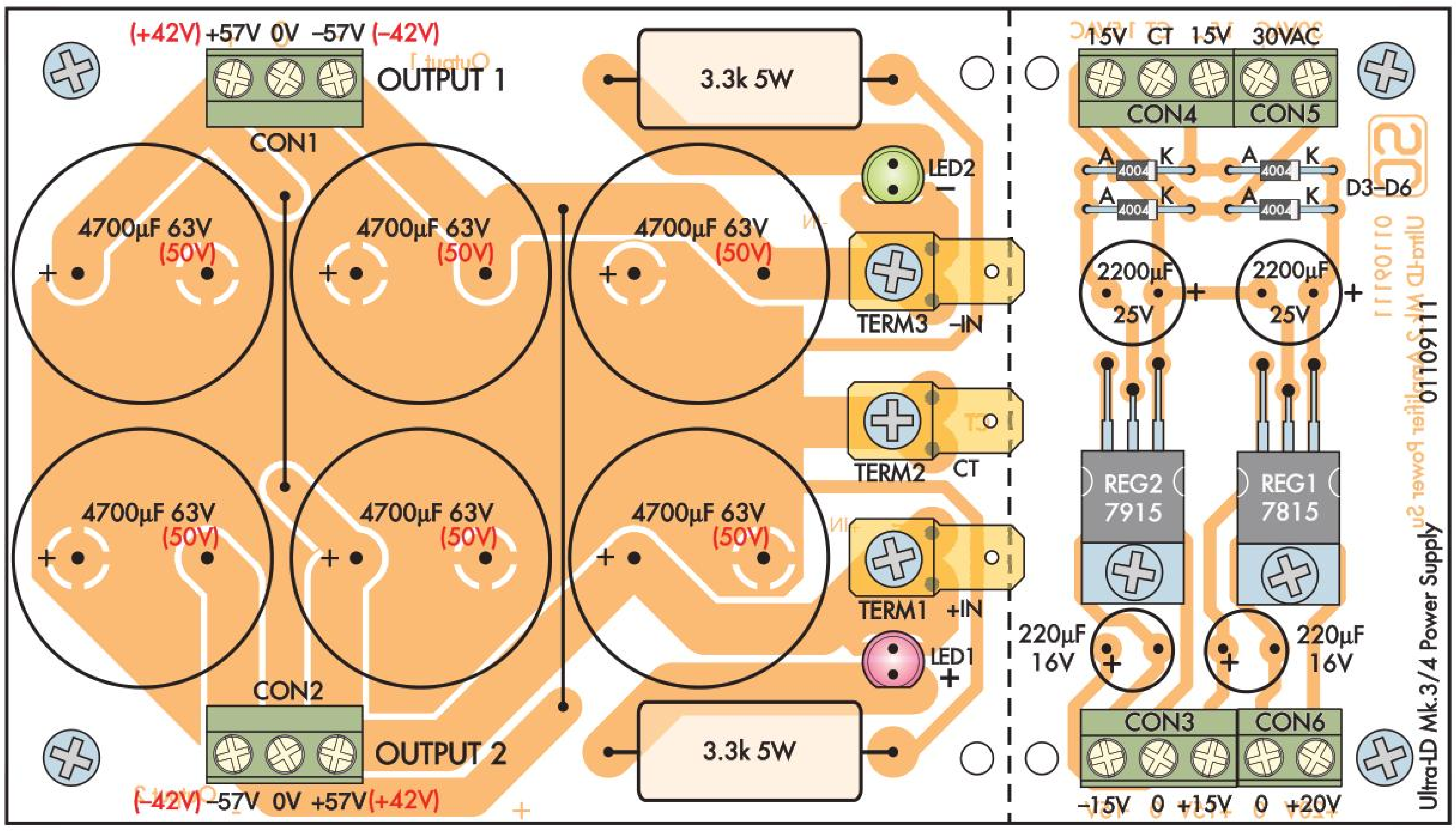
Otherwise, the Class B currents flowing through the supply leads could couple into the amplifier module(s) and ruin the performance.

Be very careful when inserting the wires into the 3-way terminal block that you get the polarity right. Refer to the wiring diagram, Fig.13, and ensure your wiring polarity matches this. The 4-way pluggable connector for CON2 is used to run a pair of heavy wires to the speaker terminal (which should ideally be twisted together) from the terminals labelled Out and GND and optionally, two more to a headphone socket, labelled HP and GND.

Initial testing

If you're confident you've built the amplifier module correctly, it is possible to simply wire it to the power supply and fire it up. But we suggest a more prudent approach,

Fig.11: use this overlay diagram to help you build the power supply PCB. You can separate the two halves and even discard the right-hand section entirely if you don't need the ±15V output. The two links shown at left should be incorporated into the top layer of the PCB if you get it from *EPE*.



so it's much safer to first wire 68Ω safety resistors in series with the supply connections as this will reduce the chance of damage if something has gone wrong.

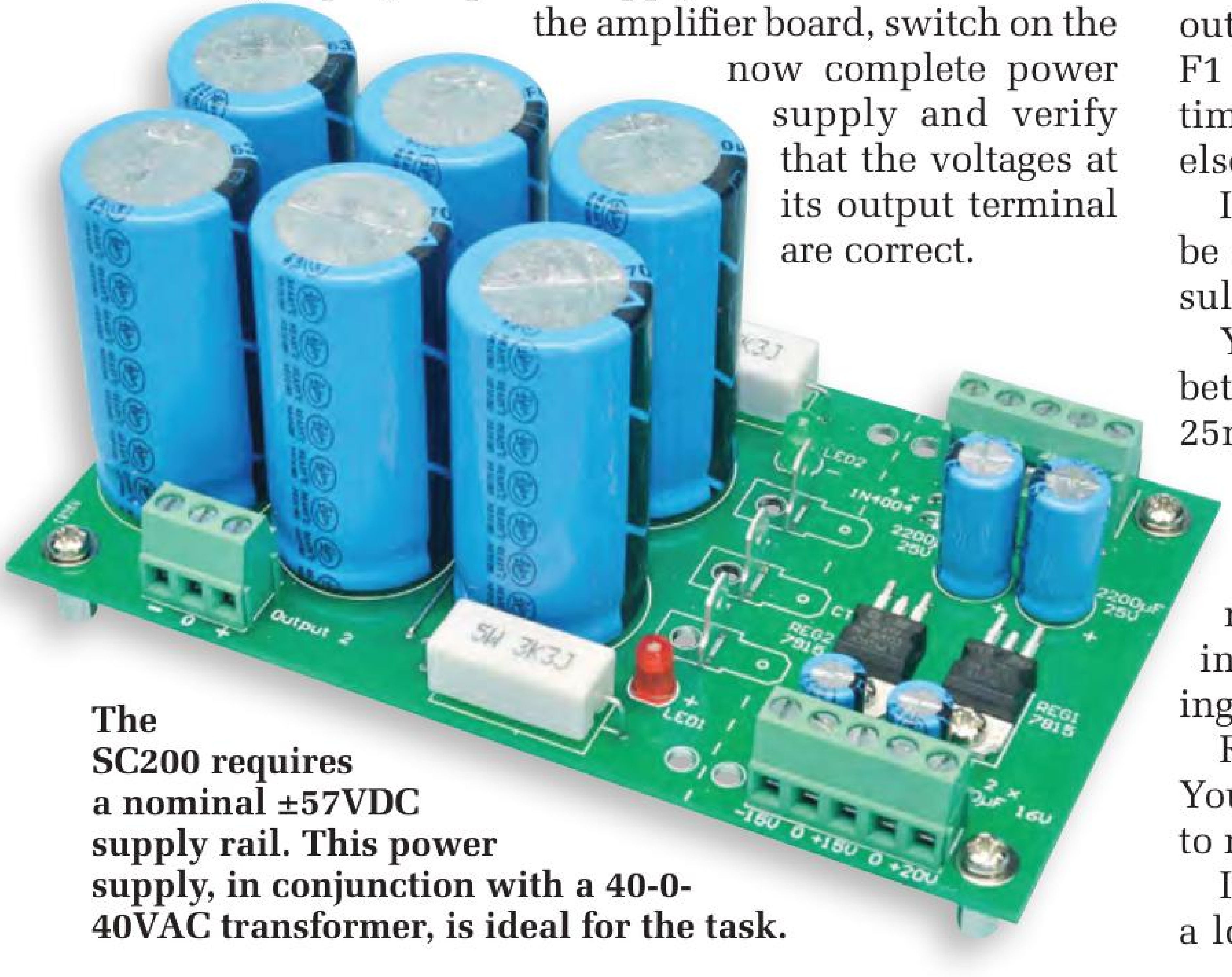
The easiest way to do this is to insert one lead of a 68Ω 5W resistor into each of the two terminals at either end of the block and do the screws up tightly, then similarly screw the other ends into a 3-way mains terminal block. You can use insulated wire or a 0.1Ω 5W resistor for the ground connection. This arrangement is shown in Fig.14.

The advantage of doing it this way is that you can easily monitor the current flowing through the resistors with a DMM (in volts mode) and the leads are unlikely to short together, as long as they are carefully arranged initially.

The other side of the terminal block is wired to the DC outputs of the power supply. This will need to be built and wired up inside an earthed case. The simplest solution is to build the power supply into the case that you intend to use for your final amplifier, and simply run an extra-long 3-way lead out of the case for testing purposes.

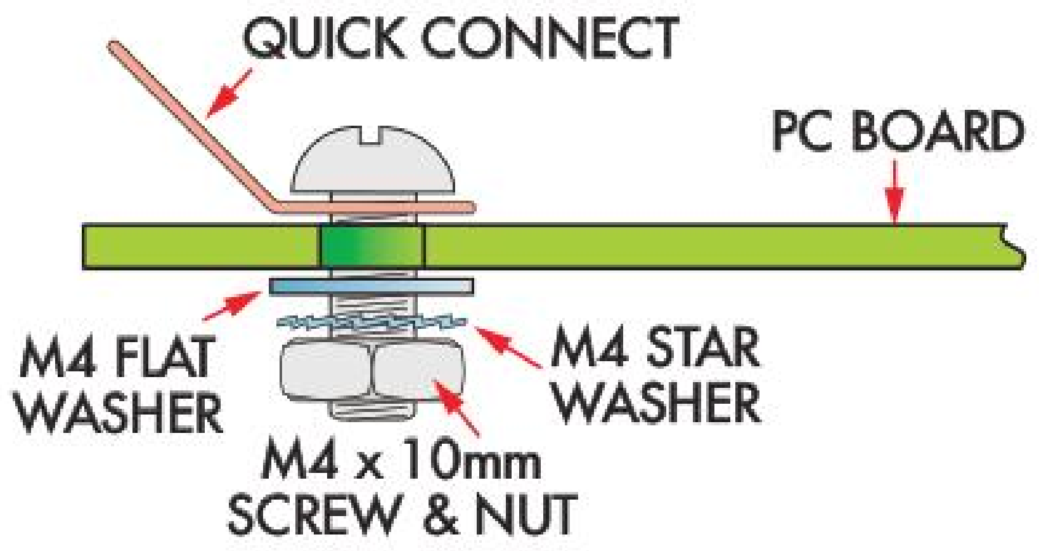
Don't skimp on this arrangement; make sure all the mains wiring is properly insulated and anchored for the tests. Once you have verified the module(s) are working you can then mount them in the case and complete the amplifier. Refer to the notes on putting the power supply together later in this article (under the 'Chassis Assembly' heading).

Before you plug the power supply connector into CON3 on the amplifier board, switch on the now complete power supply and verify that the voltages at its output terminal are correct.



The SC200 requires a nominal ±57VDC supply rail. This power supply, in conjunction with a 40-0-40VAC transformer, is ideal for the task.

Fig.12: if using the chassis-mount spade terminals on the power supply board, fit them as shown here.



The exact DC voltages will vary depending on your mains supply, but for the full power version, you should get something like 54-57V or 39-42V for the low-power version. Be especially careful to check for the correct polarity.

Switch off and wait for the LEDs on the power supply board to go out before connecting the module. Then connect a DMM set to measure volts across each safety resistor using alligator clip leads. If you don't have two DMMs, monitor one resistor. If you don't have alligator clip leads, you will have to hold the probes in place after switching power on.

Wind VR1 fully anti-clockwise and set VR2 to its halfway position using a small jeweller's screwdriver. Ensure F1 and F2 have not been fitted, then switch power on and check the onboard LEDs and the DMM readings. You should see LED1 (blue) light up along with LED 2 and 4 (red). LED6 may flicker initially, but should not stay on. Check for a reading of just under 1V across each of the safety resistors and verify that the two readings are close in value.

Assuming it's OK, switch off and wait for the LEDs to go out, which will probably take a couple of minutes. Then fit F1 and F2, switch back on and re-check everything. This time LED3 and LED5 (green) should light up but not much else should have changed.

If it does, then the output stage is suspect, eg, it could be an isolation failure on one of the output transistor insulating washers.

You can now check the output offset voltage, measuring between Out and GND on CON2. It should be less than 25mV and is usually about 10mV. Be careful not to short the two pins together!

Now rotate VR1's screw clockwise slowly while monitoring the voltage across a safety resistor. At first nothing should happen but eventually it will rise. This indicates that the Vbe multiplier is working; stop turning VR1.

Rotate VR2 and check that the offset voltage changes. You can trim it close to 0mV now, although you will need to make the final adjustment later.

If you have a scope and signal generator, you can feed a low-level signal into the amplifier (<250mV RMS) and

Parts List – SC200 Power Supply

- 1 PCB, available from the *EPE PCB Service*, coded 01109111, 141 × 80mm
- 4 3-way PCB-mount terminal blocks, 5.08mm pitch (CON1-4)
- 2 2-way PCB-mount terminal blocks, 5.08mm pitch (CON5-6)
- 3 PCB-mount or chassis-mount spade connectors
- 3 M4 × 10mm machine screws, nuts, flat washers and shake-proof washers (if using chassis-mount spade connectors)
- 4 M3 × 9mm tapped nylon spacers
- 10 M3 × 6mm machine screws
- 2 M3 shake-proof washers and nuts

Semiconductors

- 1 7815 regulator (REG1)
- 1 7915 regulator (REG2)
- 4 1N4004 1A diodes (D1-D4)
- 1 5mm green LED (LED1)
- 1 5mm yellow LED (LED2)

Capacitors

- 6 4700µF 63V [50V*] electrolytic
- 2 2200µF 25V electrolytic
- 2 220µF 16V electrolytic

Resistors

- 2 3.3kΩ 5W

Additional parts

- 1 300VA 40-0-40V + 15-0-15V transformer *OR*
- 1 160VA 30-0-30V + 15-0-15V transformer*
- 1 35A 400V chassis-mount bridge rectifier
- 1 chassis-mount IEC mains input socket with fuseholder and fuse
- Various lengths mains-rated heavy duty hookup wire
- Various spade crimp connectors Cable ties, heatshrink tubing * for lower power version

check that the output signal looks clean. Note that with the safety resistors in-circuit, it won't drive a load, nor will it handle high-swing or high-frequency signals.

Quiescent current adjustment

Switch off, wait for the LEDs to go off and remove the safety resistors. These can now be soldered across a pair of blown fuses to make handy resistor fuse adaptors; see the adjacent photo. Fit these in place of F1 and F2 and wire up the power supply direct this time, as shown in Fig.13.

Given that the earlier tests were successful, it's unlikely anything will go wrong at this stage, but it's still a good idea to have the safety resistors in place of the fuses initially. These limit the current through the output stage to about 840mA if there is a fault. Note that the 68Ω resistors will quickly burn out under such circumstances (since they would be dissipating over 40W).

Now use the following procedure to set the quiescent current and trim out the offset voltage.

STEP 1 – check that the safety resistors are installed and that their leads can't short to any adjacent parts (note: do NOT connect the loudspeaker to the amplifier during this procedure).

STEP 2 – connect a DMM set to volts across one of the safety resistors (alligator clip leads are extremely handy in this situation).

STEP 3 – turn trimpot VR1 fully anti-clockwise. This can take as many as 25 turns but it will continue to turn even so. Many (but not all) multi-turn trimpots click when they are at the end-stop. If in doubt, check the resistance across it – it should be about 1kΩ.

STEP 4 – check that the power supply is off and that the filter capacitors are discharged (LEDs off!), then connect the ±57V supply to the module. Check that the supply polarity is correct, otherwise the amplifier will be damaged when power is applied.

STEP 5 – apply power and check the voltage across the 68Ω resistor. It should be less than 1V (it may jump around a bit). If the reading is over 10V, switch off immediately and check for faults.

STEP 6 – using an insulated adjustment tool or a small flat-bladed screwdriver, slowly adjust the trimpot clockwise. Be careful not to short any adjacent components.

STEP 7 – after a few turns, the resistor voltage should stabilise and start to rise. Continue until it reads around 6V. It may drift a little but should be quite steady.

STEP 8 – switch off, wait for the capacitors to fully discharge (LEDs off) and replace the safety resistors with 6.5A fuses.

STEP 9 – connect a DMM set to volts between TP5 (to the upper left of D3) and TP7 (lower right of D3). If you have fitted PC stakes you can use alligator clip leads, otherwise you may need to get someone else to hold the probes in place while you perform the following steps.

STEP 10 – reapply power and check that the DMM reads close to 4.4mV. If necessary, readjust trimpot VR1 to bring the voltage close to this figure.

STEP 11 – now check the voltage between TP3 and TP7. The reading should be similar. Do the same check with TP4/TP7 and TP6/TP7. This verifies that all the output transistors are working and sharing the load current more or less equally.

STEP 12 – adjust VR2 until the voltage across the output pins is less than 0.5mV. This is easier to do if you screw a couple of bits of wire into the top two connections of the pluggable terminal block for CON2 and clip a DMM across it using alligator clip leads. Be extra careful not to short the output terminals together! Note that this is a trial-and-error process because you will probably find each time you remove the screwdriver from VR2, it will take several seconds for the output voltage to stabilise. You will need to make very small adjustments towards the end of the process.

It's a good idea to recheck the quiescent current (ie, between TP5 and TP7) after the amplifier has been idling for a few minutes with the lid on. If the reading is more than 5mV, readjust VR1 anti-clockwise to bring it back below this figure. The stability is such that it should stay below this figure but it's a good idea to check.

That completes the adjustments. Note, however, that if you wish to repeat the above procedure (ie, with the 68Ω resistors in place), you will first have to reset VR1 to minimum (ie, fully anti-clockwise). If you don't do this, the amplifier may latch up when power is reapplied and could burn out the safety resistors.

Troubleshooting

If there's a fault in the module, a likely symptom is either excessive voltage across the safety resistors or the amplifier output voltage is pegged near one of the ±57V supply rails.

Soldering a 5W resistor across a blown fuse makes for a handy way to limit current through the amplifier's output stage during testing and adjustment.

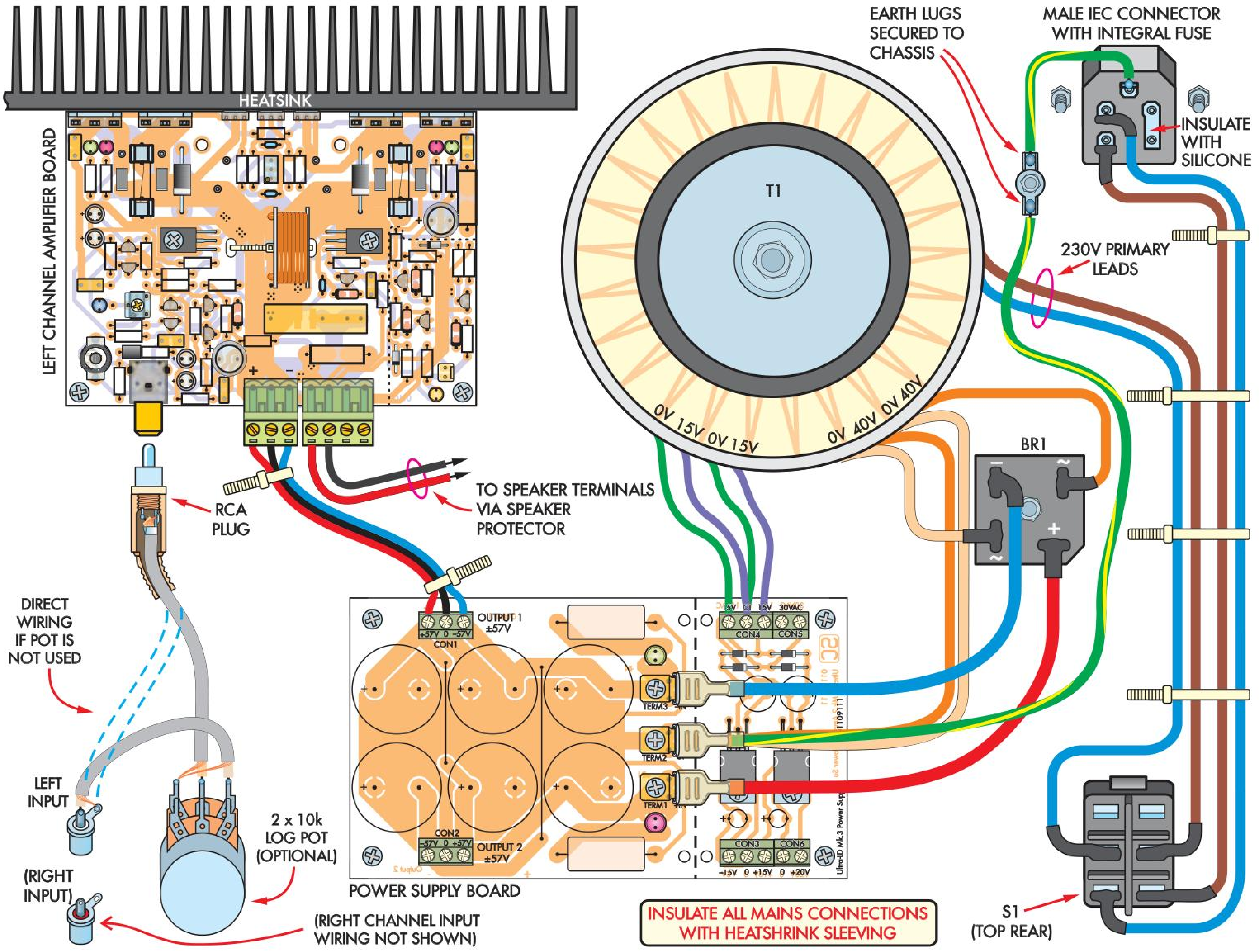

HEATSINK
LEFT CHANNEL AMPLIFIER BOARD
EARTH LUGS SECURED TO CHASSIS
MALE IEC CONNECTOR WITH INTEGRAL FUSE
INSULATE WITH SILICONE
T1
230V PRIMARY LEADS
0V 15V 0V 15V
0V 40V 0V 40V
BR1
TO SPEAKER TERMINALS VIA SPEAKER PROTECTOR
RCA PLUG
DIRECT WIRING IF POT IS NOT USED
LEFT INPUT
2 x 10k LOG POT (OPTIONAL)
(RIGHT INPUT)
OUTPUT 1 ±57V
+57V 0 –57V
CON1
CON2
–57V 0 +57V
OUTPUT 2 ±57V
15V CT 15V 30VAC
CON4
CON5
TERM3
TERM2
TERM1
CON3
CON6
–15V 0 +15V 0 +20V
POWER SUPPLY BOARD
(RIGHT CHANNEL INPUT WIRING NOT SHOWN)
INSULATE ALL MAINS CONNECTIONS WITH HEATSHRINK SLEEVING
S1 (TOP REAR)

Fig.13: this shows how to wire up the amplifier module, power supply, volume control and signal input. This should give you a working mono amplifier. You can connect a second amplifier board to the same power supply, in a similar fashion as shown here, to build a stereo amplifier. Don't forget the speaker protector!

If this happens, switch off and wait for the power supply capacitors to discharge. Then check that all the large transistors are properly isolated from the heatsink. You should also carefully inspect all the solder joints on the underside of the board, to make sure that they all have good, shiny fillets and also check to make sure that all the correct component types and values are in the intended locations and none of the polarised components have been installed backwards.

If you still can't find the fault, you will need to power the amplifier up without fuses or safety resistors fitted. Then check the various voltages shown in the circuit diagram, Fig.1 on pages **20** and **21** of the January issue, with reference to the overlay diagrams of Fig.4 on page **26** in the February issue. If you find a voltage which is clearly wrong, this may give you a clue to where the fault lies.

Chassis assembly

If you want to build a complete stereo SC200 amplifier, it's vital to include a loudspeaker protection module. For a suitable module, see our design in the December 2016 issue.

Volume control

Once you've built the power supply, amplifier module(s) and speaker protector and wired them up, if you are not fitting a full preamplifier in the case, you will probably want to fit a volume control. This is quite simple and Fig.13 shows how to do it using a 10kΩ dual-gang logarithmic potentiometer.

Basically, you just need to connect the incoming signal wire to the clockwise end of the potentiometer with its shield ground to the anti-clockwise end. The reduced amplitude signals then appear at the wipers and these are connected to the signal wire for the cables going to the amplifier modules, with the shield grounds soldered together with the shields from the incoming wire (ie, to the anti-clockwise end of the potentiometer track).

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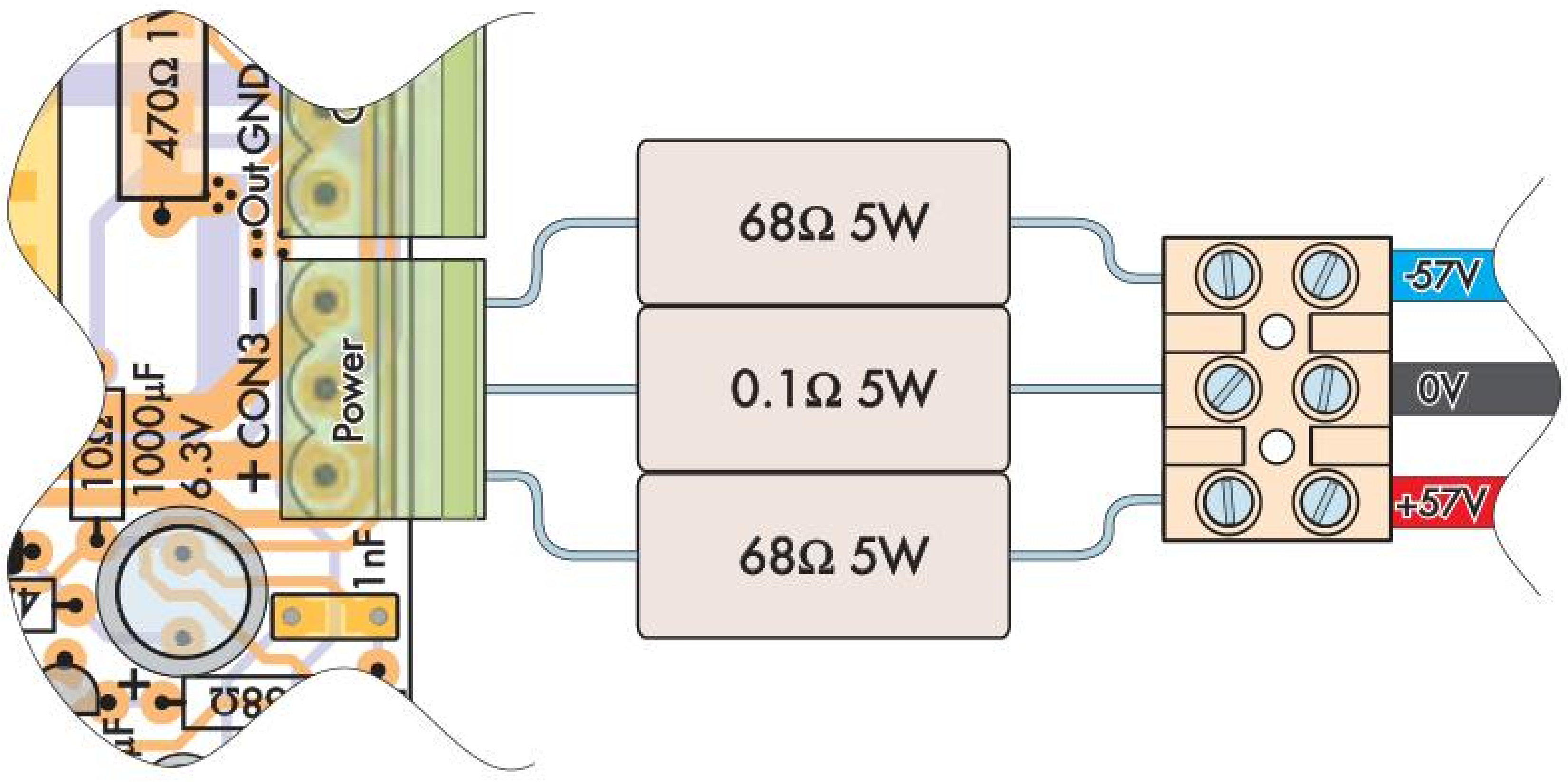
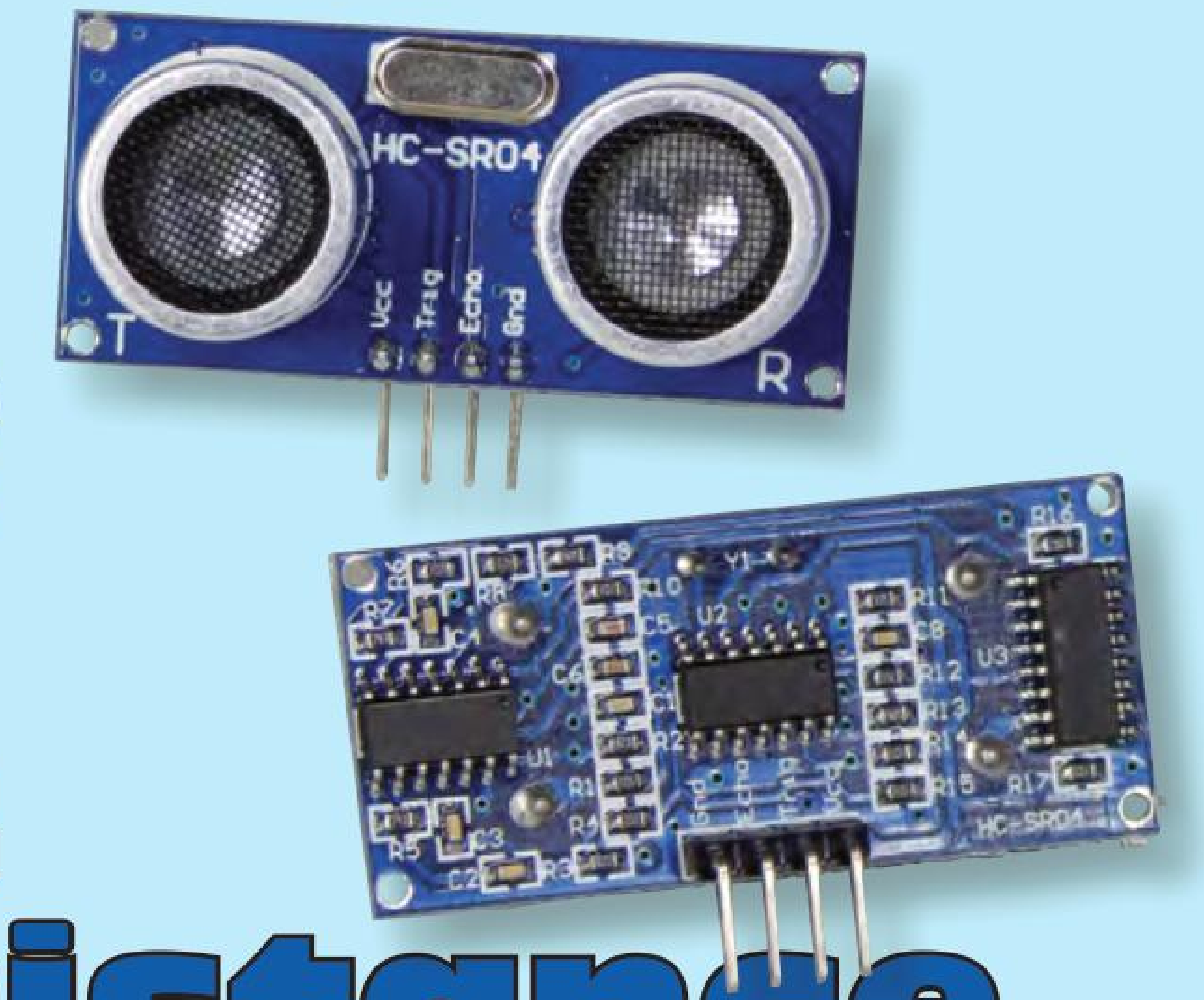

470Ω 1W
Out GND
CON3
1000µF
10Ω
Power
1nF
68Ω 5W
0.1Ω 5W
68Ω 5W
-57V
0V
+57V

Fig.14: we recommend you connect the power supply to the amplifier board as shown here the first time you power it up. This way, if there's a fault, it's much less likely to cause any damage to the module before you have time to switch the power off.

Using Cheap Asian Electronic Modules Part 2



The HC-SR04 Ultrasonic Distance Sensor Module

In the second article on cheap pre-built electronics modules, we're focusing on the HC-SR04 ultrasonic distance sensor module. We describe how the module works and show how it can be used as a hallway monitor or door sentry.

By JIM ROWE

IF THE HC-SR04 module shown in the picture looks familiar, that's because it has already been used in Geoff Graham's *Ultrasonic Garage Parking Assistant*, published in the June 2017 issue. But this module doesn't have to be used with a microprocessor module like a Micromite or an Arduino, it can also be used with much simpler circuitry.

Before we get to how it works, we should note that these ultrasonic sensor modules have been around for about six years, beginning life as an add-on 'shield' for the Arduino. Since then, they have gone through a number of iterations, all bearing the same HC-SR04 label but with various minor circuit and component changes. We suspect this has been due to various manufacturers working out ways of reducing costs, rather than seeking to achieve better performance.

The bottom line is that although some of these slightly different HC-SR04 modules are still being sold, they all seem to function and perform much the same. So don't worry if the module you buy looks a little different from that shown in the photos. The odds are that if your module carries the label HC-SR04, it will work just like any other HC-SR04.

Current HC-SR04 modules are based on a PCB measuring 45 × 20mm. On the top side of the PCB is a pair of small (16mm diameter) ultrasonic transducers with a 4MHz crystal between them.

All the components on the other side of the PCB are surface-mount types, apart from the 4-pin right-angle header at bottom centre. Fig.1 shows how it's used. It sends out a burst of ultrasonic energy from the transmitter transducer (the one marked T, on the left) and then listens via the other receiver transducer (marked R, on the right) for any echo that may be reflected back from an object in front of the module (see Fig.1).

If it detects this ultrasonic echo, it produces an output pulse with a width approximately proportional to the distance between the module's sensors and the object producing the echo.

The ultrasonic frequency used is very close to 40kHz, roughly double

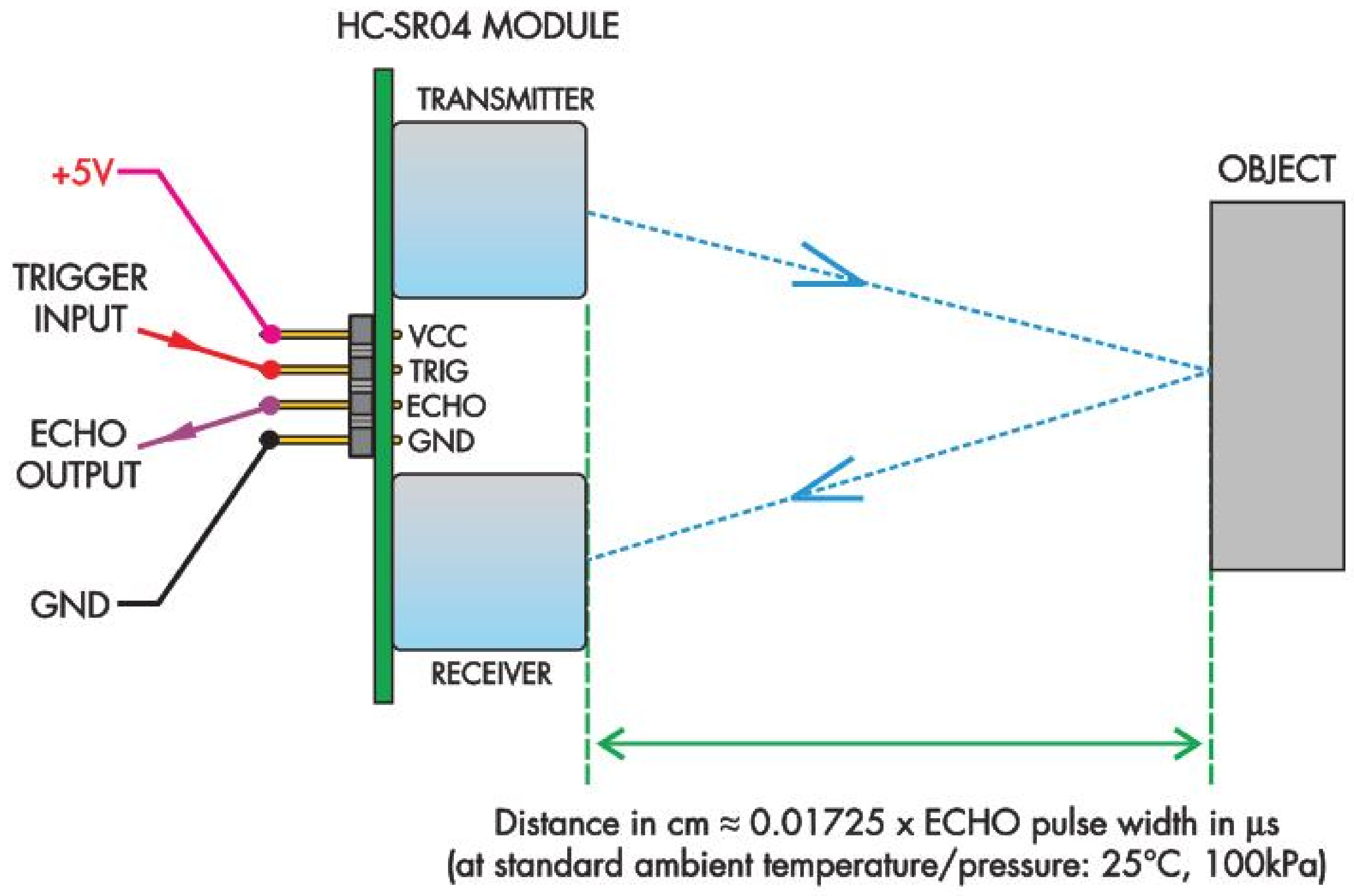

HC-SR04 MODULE
TRANSMITTER
+5V
TRIGGER INPUT
ECHO OUTPUT
GND
VCC
TRIG
ECHO
GND
OBJECT
RECEIVER
Distance in cm ≈ 0.01725 x ECHO pulse width in µs
(at standard ambient temperature/pressure: 25°C, 100kPa)

Fig.1: one ultrasonic burst is sent out from the transmitter transducer. The receiver transducer will detect this burst if it is reflected off an object in front of the module. Once detected by the receiver, an output pulse is produced with a width in microseconds of (distance in cm) ÷ 0.01725.

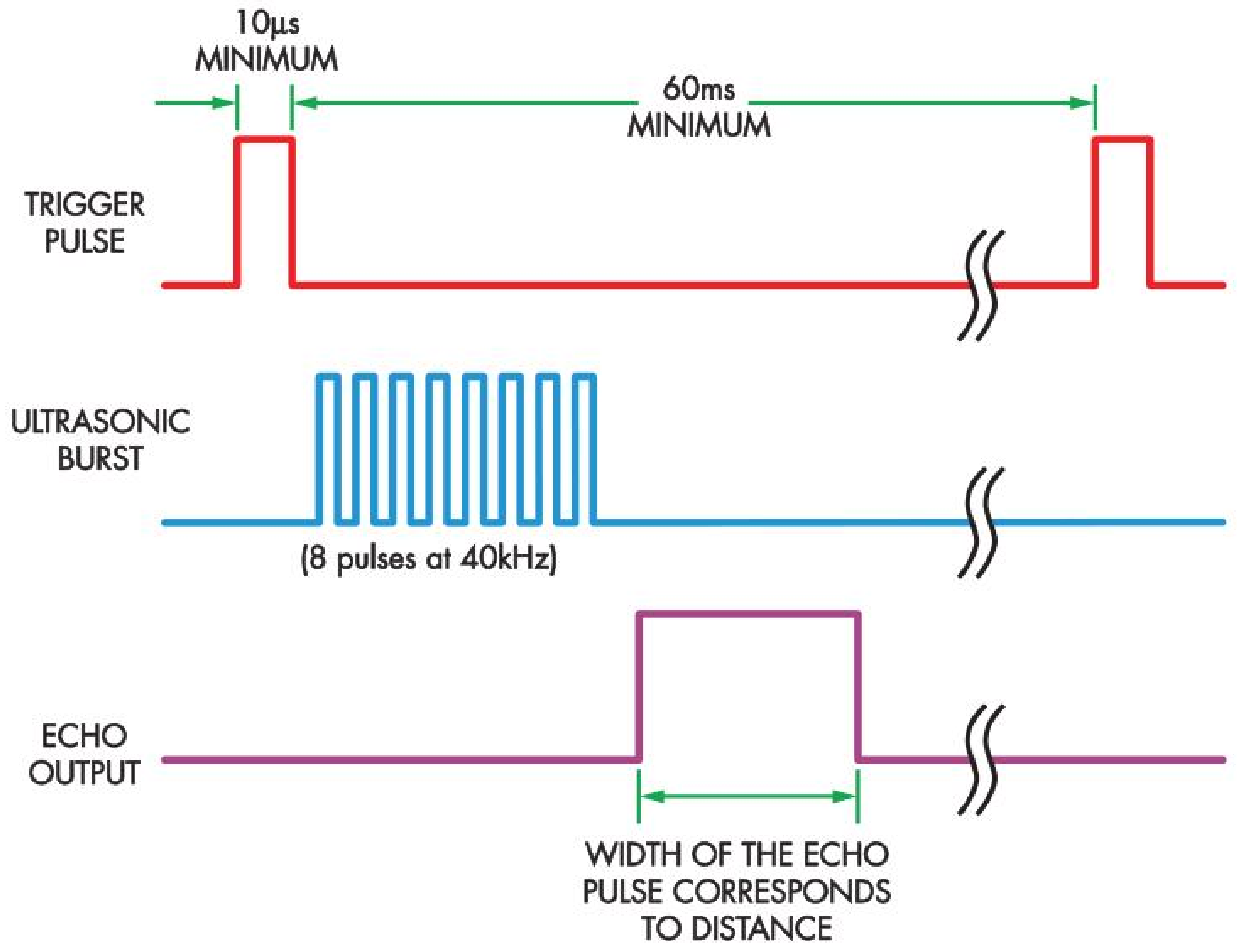

10µs MINIMUM
60ms MINIMUM
TRIGGER PULSE
ULTRASONIC BURST
(8 pulses at 40kHz)
ECHO OUTPUT
WIDTH OF THE ECHO PULSE CORRESPONDS TO DISTANCE

Fig.2: there must be a delay of 60ms between trigger pulses to prevent late echoes from affecting successive readings.

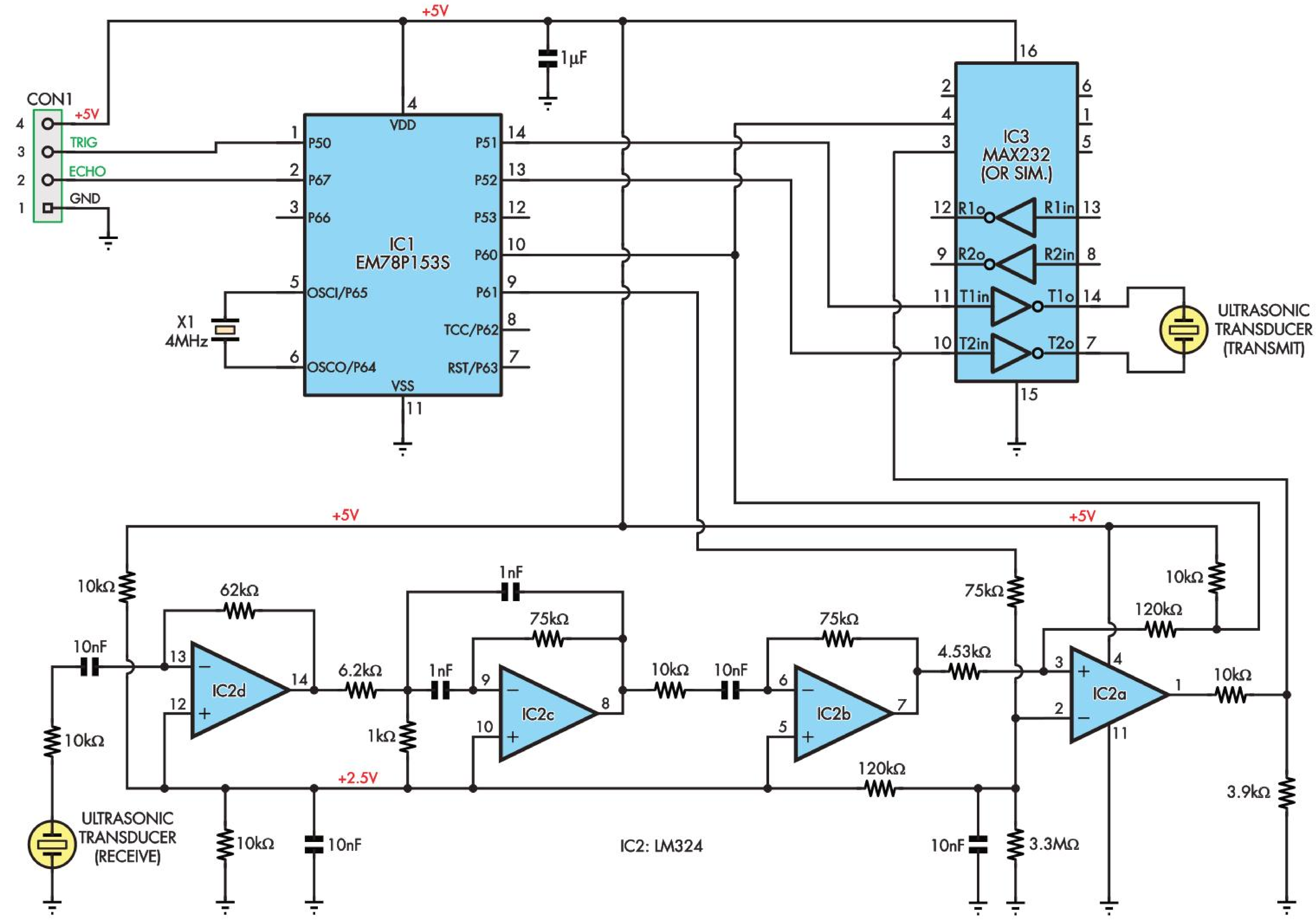

+5V
1µF
CON1
4
3
2
1
+5V
TRIG
ECHO
GND
VDD
P50
P67
P66
P51
P52
P53
P60
P61
IC1
EM78P153S
OSCI/P65
OSCO/P64
TCC/P62
RST/P63
VSS
X1
4MHz
16
15
IC3
MAX232
(OR SIM.)
R1o
R1in
R2o
R2in
T1in
T1o
T2in
T2o
ULTRASONIC
TRANSDUCER
(TRANSMIT)
+5V
+5V
10kΩ
62kΩ
1nF
75kΩ
75kΩ
75kΩ
10kΩ
120kΩ
10nF
IC2d
6.2kΩ
1nF
IC2c
10kΩ
10nF
IC2b
4.53kΩ
IC2a
10kΩ
1kΩ
10kΩ
+2.5V
120kΩ
3.9kΩ
ULTRASONIC
TRANSDUCER
(RECEIVE)
10kΩ
10nF
IC2: LM324
10nF
3.3MΩ

Fig.3: complete circuit diagram for the HC-SR04 ultrasonic sensor module. When IC1 detects a TRIG pulse at pin 1, a 40kHz burst signal of eight pulses is generated at pins 13 and 14 of IC1. This is taken to pins 10 and 11 of IC3 respectively, and output at pins 7 and 14 connecting to the transmit transducer.

the highest frequency that can be heard by human ears. The burst of transmitted energy consists of eight pulses at 40kHz, so the transmitted burst lasts for only 200µs, as shown in Fig.2.

Since the speed of sound in air at 25°C and 100kPa (ie, 1 bar) is close to 345m/s (= 0.0345cm/µs) and the distance travelled by the ultrasonic burst energy corresponds to double the distance between the transducers and the reflecting object, we can calculate the distance from the delay as follows:

$$\textbf{distance in cm} = \frac{\textbf{0.0345 x echo pulse width (µs)}}{\textbf{2}}$$

$$= \textbf{0.01725 x echo pulse width (µs)}$$

As shown in Fig.2, each measurement cycle begins when a positive trigger pulse of at least 10µs duration is applied to the HC-SR04 module's trigger input pin. When the echo has been detected, it then produces a pulse at the echo output pin. Note that there should be at least 60ms between trigger pulses, to prevent late echoes from one cycle causing false readings on the next. So in practice, it's a good idea to limit the trigger pulse frequency to no more than 16Hz.

Circuit details

The full circuit for the HC-SR04 module is shown in Fig.3. It is based on an EM78P153S microcontroller (IC1), a low-power 8-bit CMOS device made by Elan Microelectronics in Hsinchu, Taiwan. This device has a 1024 × 13 bits one-time programmable (OTP) ROM plus 32 bytes of on-chip SRAM, and comes in a 14-pin SOIC package. It runs here with a 4MHz crystal between pins 5 and 6.

When a TRIG pulse arrives at pin 1 of IC1 (from pin 3 of CON1), the controller generates a 40kHz burst signal of eight pulses at pins 13 and 14, with one pin 180° out of phase with the other. These go to pins 10 and 11 of IC3, a bus driver IC very similar to the MAX232. The outputs from IC3 (pins 7 and 14) connect across the transmitter transducer, effectively driving it in bridge mode to emit the bursts of ultrasonic energy.

Echoes picked up by the receive transducer pass through the four sections of IC2, an LM324 quad op amp. These provide amplification, bandpass filtering and phase detection, with the result that a received echo pulse is fed back to pin 10 of IC1. The micro then compares the timing of the leading edge of this received echo pulse with the leading edge of the transmitted burst fed to IC3 and the transmit transducer, and produces an echo output pulse at pin 2 with its width equal to the time difference. This echo output pulse appears at pin 2 of CON1.

How it's used

If you want to use the HC-SR04 module to actually measure the distance to an object or wall in front of it, the best way to do it is to hook it up to a microprocessor module like an Arduino, Micromite or Raspberry Pi. The micro's program generates the trigger pulse to the HC-SR04, then measures the length of the echo pulse and calculates the corresponding distance.

There's no need to worry about writing a program to do these tasks for you, because many people have already produced programs to do this. A quick search on the Arduino website (**www.arduino.cc**) or by using Google will find a sample program for the micro you're using in short order.

If you want to use the HC-SR04 with a Micromite, Geoff Graham has already built a DISTANCE function into his MMBasic programming language for the Micromite family to make it really easy.

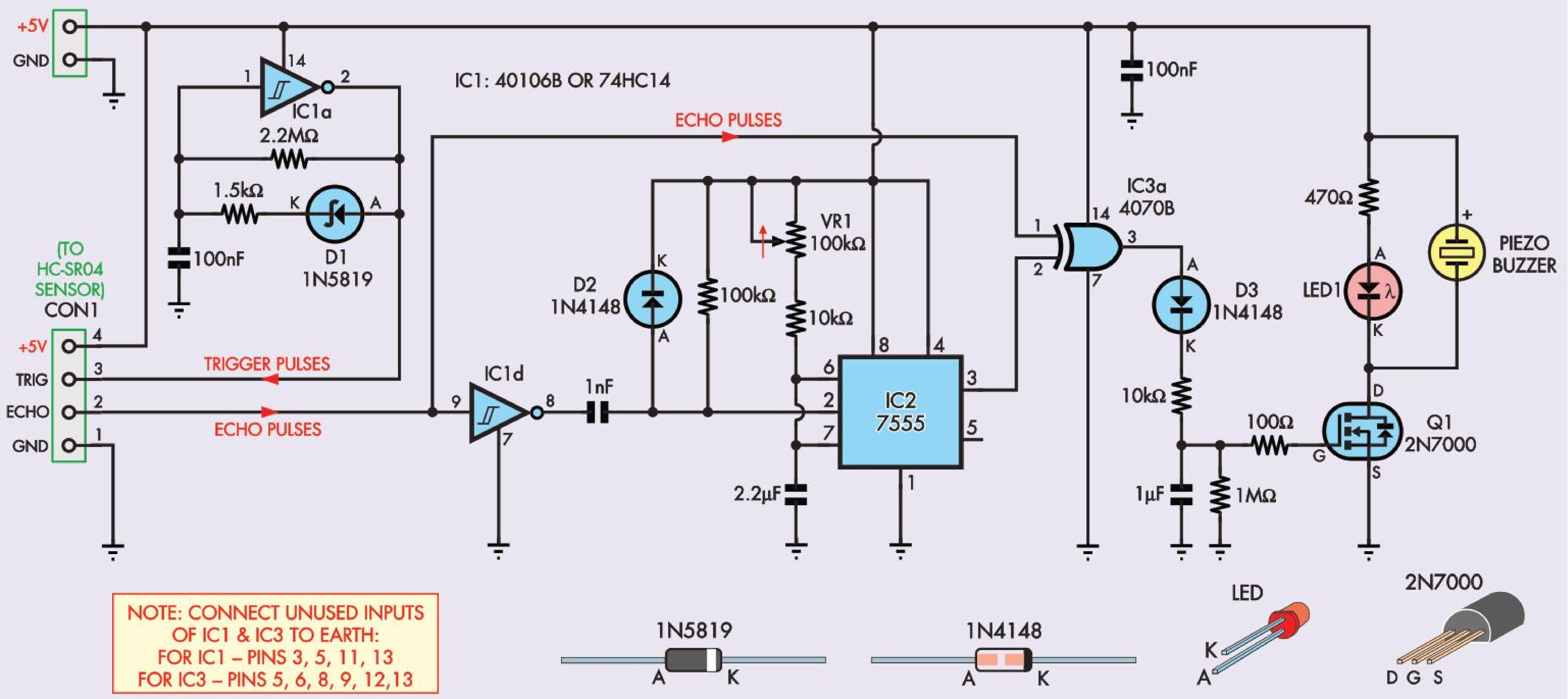


Fig.4: complete circuit for an ultrasonic intruder alarm using an HC-SR04 module. IC1a generates 60µs-wide trigger pulses at 12Hz, which are fed to pin 3 of CON1. The echo pulses trigger monostable multivibrator IC2, and IC3a then compares the width of the resulting pulse to the echo pulse. If these differ, LED1 lights and the piezo buzzer sounds.

All you have to do to get the Micromite to trigger the HC-SR04 and then calculate the object distance from the echo pulse is use this one-line function call:

```
d = DISTANCE(trig, echo)
```

Where ‘d’ is the distance in centimetres, ‘trig’ is the Micromite’s I/O pin connected to the HC-SR04’s trigger input pin and ‘echo’ is the I/O pin connected to the HC-SR04’s echo output pin.

The only extra step is to connect the HC-SR04’s +5V and GND pins to the corresponding pins of your Micromite.

If you want to display the result ‘d’ on an alphanumeric LCD, you can do this using commands like:

```
LCD INIT …
LCD 1, 2, “Distance = “
LCD 2, 6, STR$(d)
```

and so on.

You can get a good idea of what’s involved in using the HC-SR04 with a Micromite from Geoff Graham’s article describing the *Ultrasonic Garage Parking Assistant*.

But say you want to use this module without a microcontroller at all. That’s fairly straightforward, as we’ll now demonstrate.

A simple intruder alarm

For example, to use it as an ultrasonic intruder alarm, have a look at the circuit shown in Fig.4. It uses three low-cost CMOS ICs, a 2N7000 MOSFET, three diodes, one LED, a piezo buzzer and some passive components. This circuit and the HC-SR04 operate from a common 5V DC power supply, which can be from a USB plugpack or USB power bank.

IC1 is a hex Schmitt trigger inverter package and we’re using just two sections of it, IC1a and IC1b. IC1a at upper left is connected as a relaxation oscillator, to generate a stream of 60µs-wide pulses at a frequency of about 12Hz, ie, with a pulse spacing of about 83ms. These form the trigger pulses, which are fed to the HC-SR04 via pin 3 of CON1.

The rest of the circuit monitors the width of the echo pulses sent back from the HC-SR04 via pin 2 of CON1. If this varies significantly (indicating that something has moved between the sensor and the nearest object, like the opposite wall of your entry hall), it sounds the alarm by switching on LED1 and the piezo buzzer connected across it.

This section is a little more complex. First, the incoming echo pulse passes through inverter IC1d, so that its leading edge is negative-going. The 1nF capacitor and 100kΩ resistor then form a differentiator circuit, which develops a narrow negative-going pulse from the negative-going leading edge of the inverted pulse.

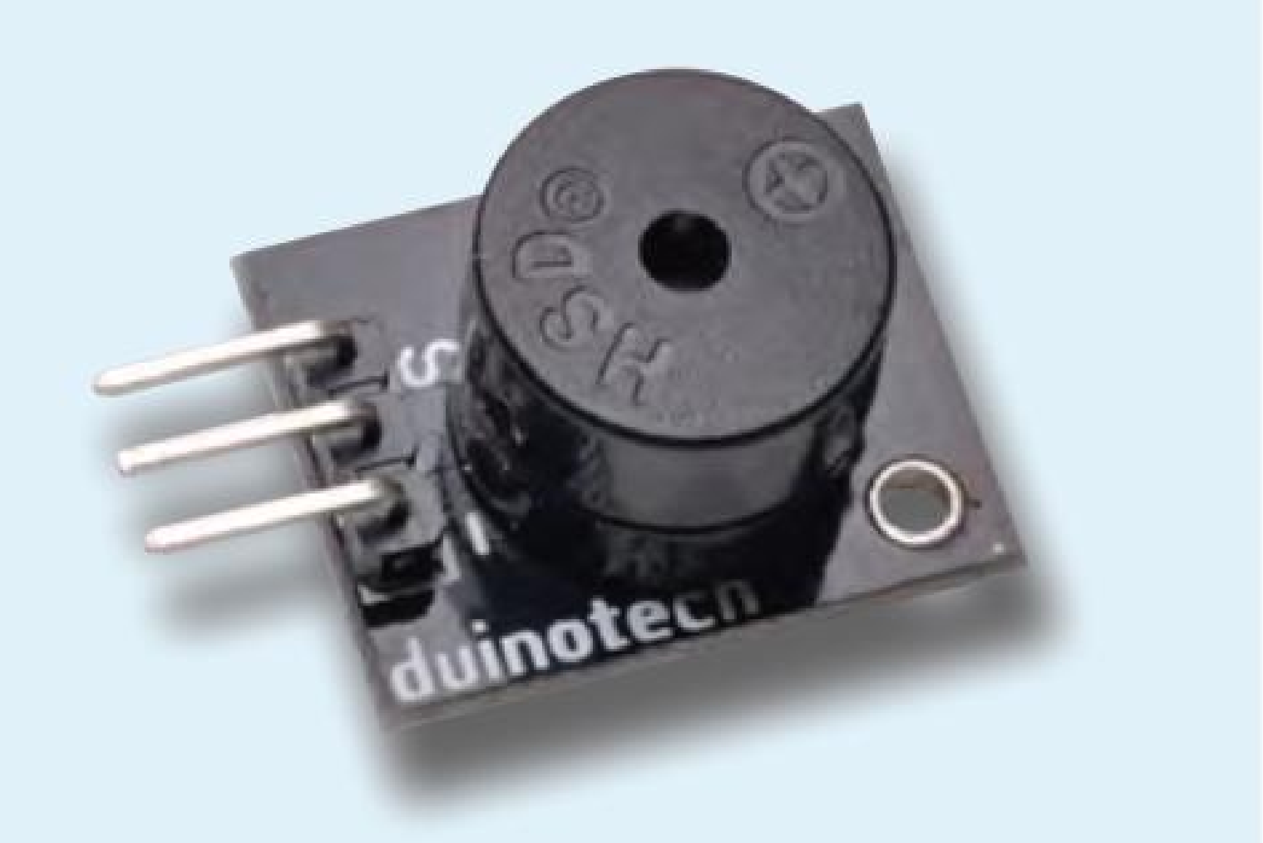
This is then used to trigger IC2, a 7555 CMOS timer chip connected as a one-shot multivibrator. When IC2 is triggered, its output pin 3 switches high for a short time, determined by the 2.2µF capacitor connected from pins 6 and 7 to ground and the resistance connected between the same two pins and the +5V line.

As shown, this resistance is the series combination of a 10kΩ resistor and VR1, a 100kΩ pot. So by varying VR1, we can vary the width of the pulse generated each time the one-shot is triggered.

The output of IC2 is connected to pin 2 of IC3a, one section of a 4070B quad XOR (exclusive-OR) gate. The echo pulses from the HC-SR04 are fed to pin 1 of IC3, the second input of the same XOR gate. Since the output of an XOR gate is high only when one of its inputs is high and the other low, it forms a pulse-width comparator.

Consider the situation where the HC-SR04 sensor is facing a wall, say 1.5m or 150cm away. The echo pulses fed back from the sensor will be very close to 8.7ms wide and these are fed to input pin 1 of IC3a.

If we adjust VR1 so that IC2 also produces 8.7ms-wide pulses, then since they start at virtually the same instant at the start of the echo pulse, both inputs of XOR gate IC3a will rise and fall at the same time. As a result, the output of IC3a (pin 3) will remain low at all times.



This tiny active piezo transducer module from Jaycar can be used in the intruder alarm instead of the piezo buzzer.

But if someone moves in front of the HC-SR04, this will cause the echo pulses to shorten, because the ultrasonic energy reflected back by the person or object will be travelling over a smaller distance. So the echo pulse width will drop briefly to say 5-6ms, and as a result the inputs of IC3a will no longer be synchronised.

Although the pulses fed to pin 2 will still be high for 8.7ms, the echo pulses being fed to pin 1 will drop low after 5-6ms, so the output of IC3a will switch high for the remaining 2.7-3.7ms. These positive-going pulses will very quickly charge up the 1µF capacitor in the gate circuit of MOSFET Q1, via diode D3 and the 10kΩ series resistor, and this will turn on Q1, causing LED1 to light and the piezo buzzer to sound the alarm.

Then, when the intruding person or object moves away again and the echo pulses return to their original width of 8.7ms, the pulses fed to the two inputs of IC3a will be again be synchronised. There will be no more output pulses from IC3a and the 1µF capacitor will be discharged by the 1MΩ resistor connected across it. So within a couple of seconds, the buzzer and LED will switch off.

The circuit is quite easy to set up, too. All you need to do is wire it up and connect it to the HC-SR04 module using a suitable length of 4-conductor cable. Then mount the sensor module on one side of the hall or doorway you want to monitor, facing either a wall or a large fixed object such as a dresser, a chest of drawers or a filing cabinet.

Next, set pot VR1 to its fully anticlockwise (ie, minimum resistance) position and turn on the 5V power supply. You'll find that LED1 will immediately light, and if you have a piezo buzzer connected as well, it will sound. That's because the pulses being generated by the one-shot IC2 will be shorter than the echo pulses coming from the HC-SR04.

Now slowly turn pot VR1 clockwise until LED1 turns off and the piezo buzzer goes silent. Your intruder alarm will then be set up and ready to detect the presence of a 'foreign body' in the space between the sensor and its reflecting wall. So we've done all this without a microprocessor – apart from the EM78P153S micro inside the HC-SR04 sensor module itself, of course.

Parts List

1 HC-SR04 ultrasonic sensor
1 active piezo transducer module ***OR***
1 piezo buzzer
1 100kΩ trimpot (VR1)

Semiconductors

1 1N5819 diode (D1)
2 1N4148 diodes (D2)
1 LED, any colour (LED1)
1 2N7000 MOSFET (Q1)
1 40106B or 74HC14 CMOS IC (IC1)
1 LM7555 CMOS timer IC (IC2)
1 4070B quad XOR gate IC (IC3)

Capacitors (16V)

1 2.2µF
1 1µF
2 100nF
1 1nF

Resistors (0.25W, 5%)

1 2.2MΩ
1 1MΩ
1 100kΩ
2 10kΩ
1 1.5kΩ
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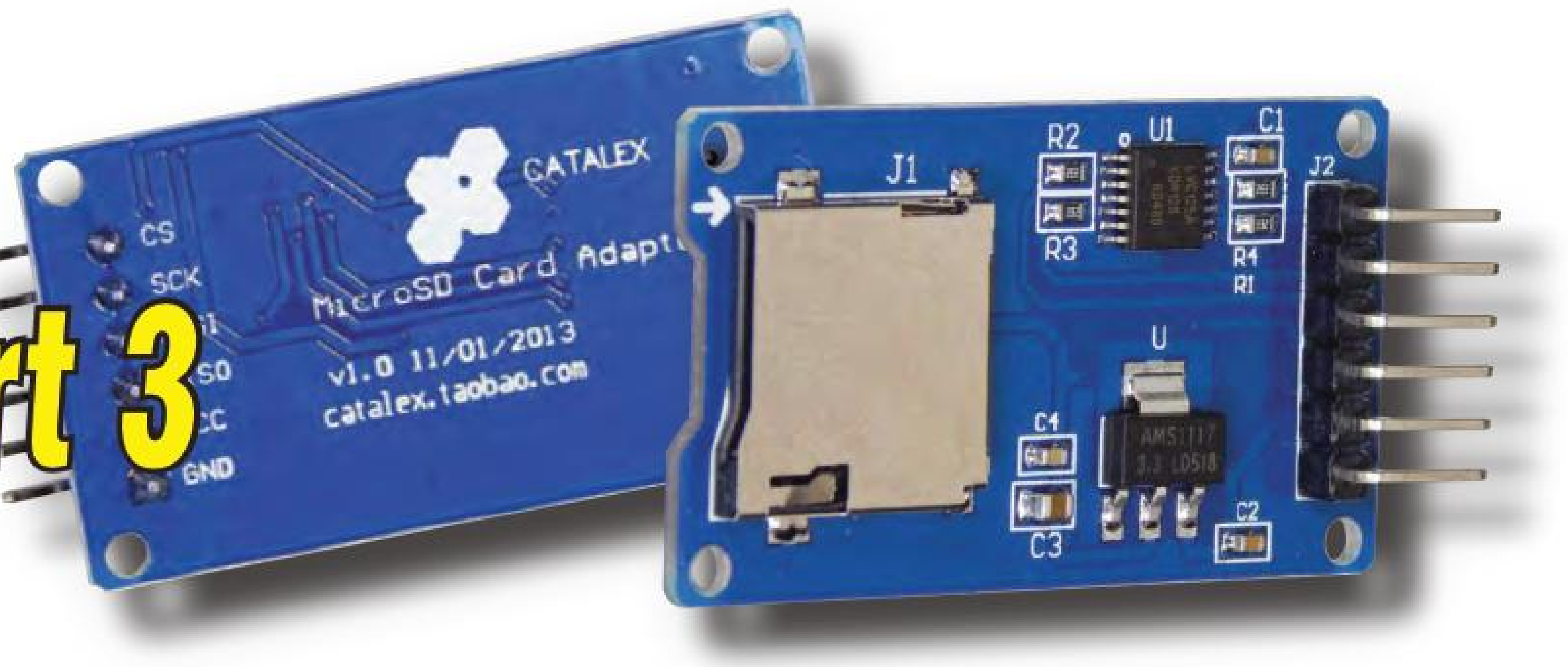
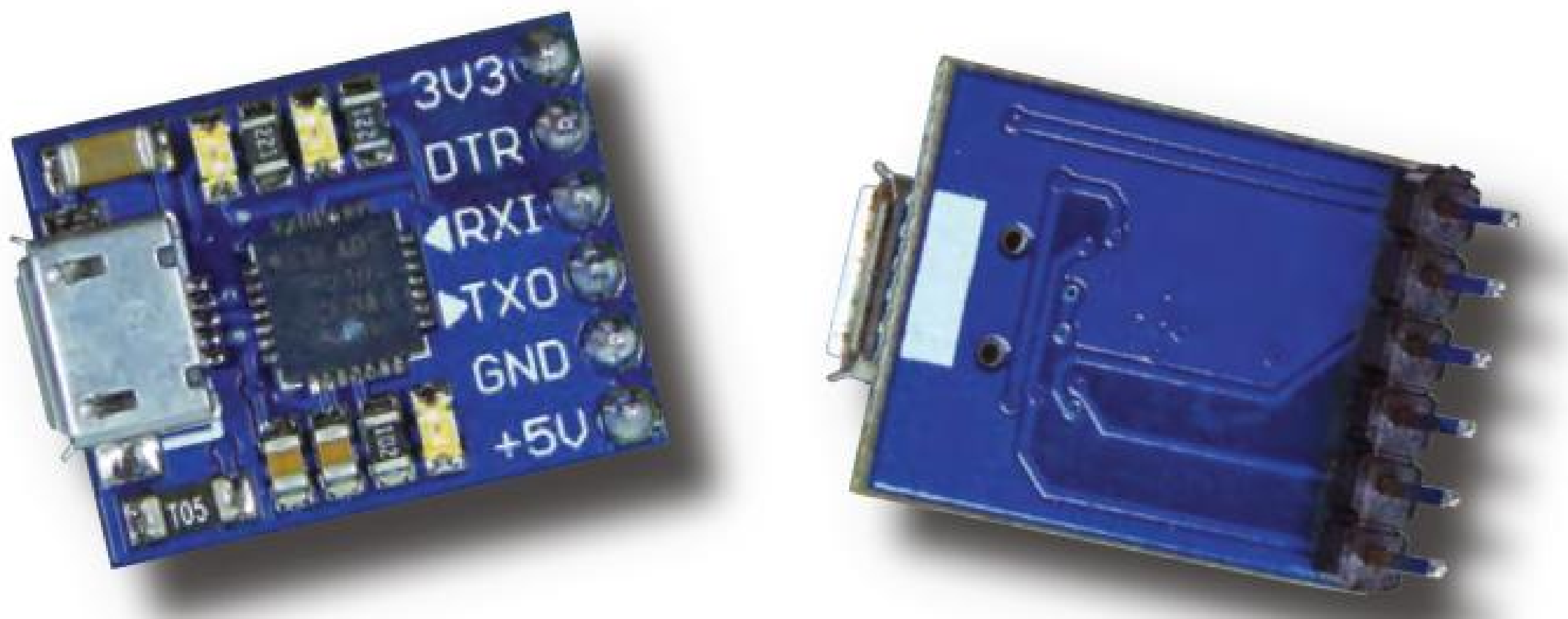
Using Cheap Asian Electronic Modules Part 3





Computer Interface Modules

Want to connect a microcontroller to your PC? How about interfacing with a microSD memory card? These low-cost modules make life really easy! Jim Rowe shows you how.

THE SECOND MODULE we're looking at this month has been used in a number of recent projects. It's a serial USB-UART (universal asynchronous receiver/transmitter) bridge which allows just about any microcomputer or peripheral module to exchange data with a PC, via a standard USB port.

Let's start by explaining what is meant by the rather clumsy term'serial USB-UART bridge'. First, a UART is an interface which can operate in one of several different common serial protocols. The serial protocol we're most interested in (and which is most widely used) is 3.3V 'TTL' RS-232. The term 'bridge' simply refers to the fact that this module allows data to pass between the USB interface and UART interface unchanged.

In fact, we've already described a device with essentially the same purpose, the Microchip MCP2200 'protocol converter' used in the *USB/RS-232C Serial Interface* which was published in the April 2015 issue.

Note that for a UART to provide a fully compatible RS-232 serial port, as used in many now-obsolete PCs, it's necessary to provide level shifting from the UART's 3.3V (TTL) signalling levels to the RS-232 bipolar logic levels of ±3-15V.

But these days, RS-232 is commonly used for short-range communications between microcontrollers and bridges and in this case, the TTL signal levels are all you really need.

The first serial USB-UART bridge modules to become popular were based around British firm FTDI's improved FT232R converter chips. However, these chips became so popular that some Asian firms made 'clones' of them, even going so far as copying the package markings.

Understandably, this upset FTDI and as a result they released a new version of their Windows VCP driver which was able to identify when a clone chip was being used and disable it. This 'clone killer' driver was included in an automatic update that Microsoft unwittingly provided to Windows users.

As a result, thousands of people found that their low-cost USB-UART converter modules, some inside commercial products, suddenly stopped working and became worthless. Naturally, this made many people cautious of buying any converter based on the FTDI FT232R chip, because of the difficulty in ensuring that you are buying a genuine FTDI chip rather than a clone chip that would stop working as soon as you tried to use it with Windows.

As a result of this, CP2102-based USB-UART bridges have become very popular. These are not only less expensive than FT232-based modules but are (currently) free from such driver issues.

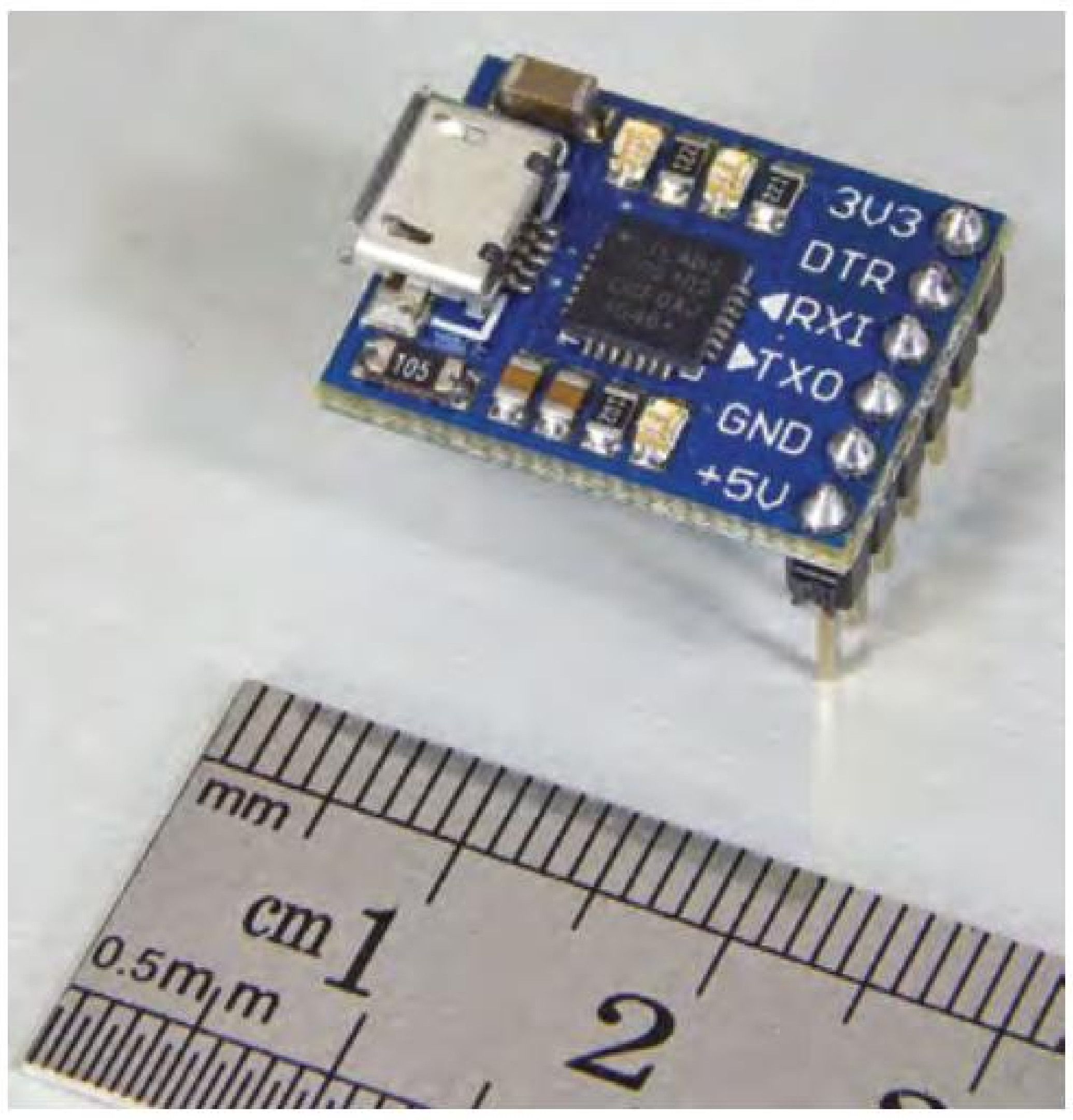
A good example of this type of module is the tiny one shown in the photo to the right. This same module has been used in recent projects and can be used with virtually any Micromite to program the micro as well as debug the software or load data into or out of the micro's RAM.

The CP2102-based bridge

As you can see from the photo and circuit diagram in Fig.1, there's very little in this module apart from the CP2102 chip (IC1), three indicator LEDs and half a dozen passive components.

The internals of IC1's tiny (5 × 5mm) 28-pin QFN SMD package are shown in the internal block diagram, Fig.2. It's conceptually quite simple but involves tens of thousands of logic gates and memory cells as well as carefully designed analogue circuitry.

The main functional blocks are the USB transceiver at lower left, the USB function controller at lower centre and the UART block at lower right with its full range of data and handshaking inputs and outputs. Notice that there's also an internal 1024-byte EEPROM



A CP2102 module, measuring only 20 × 16mm. Two of the indicator LEDs glow when data is being transmitted.

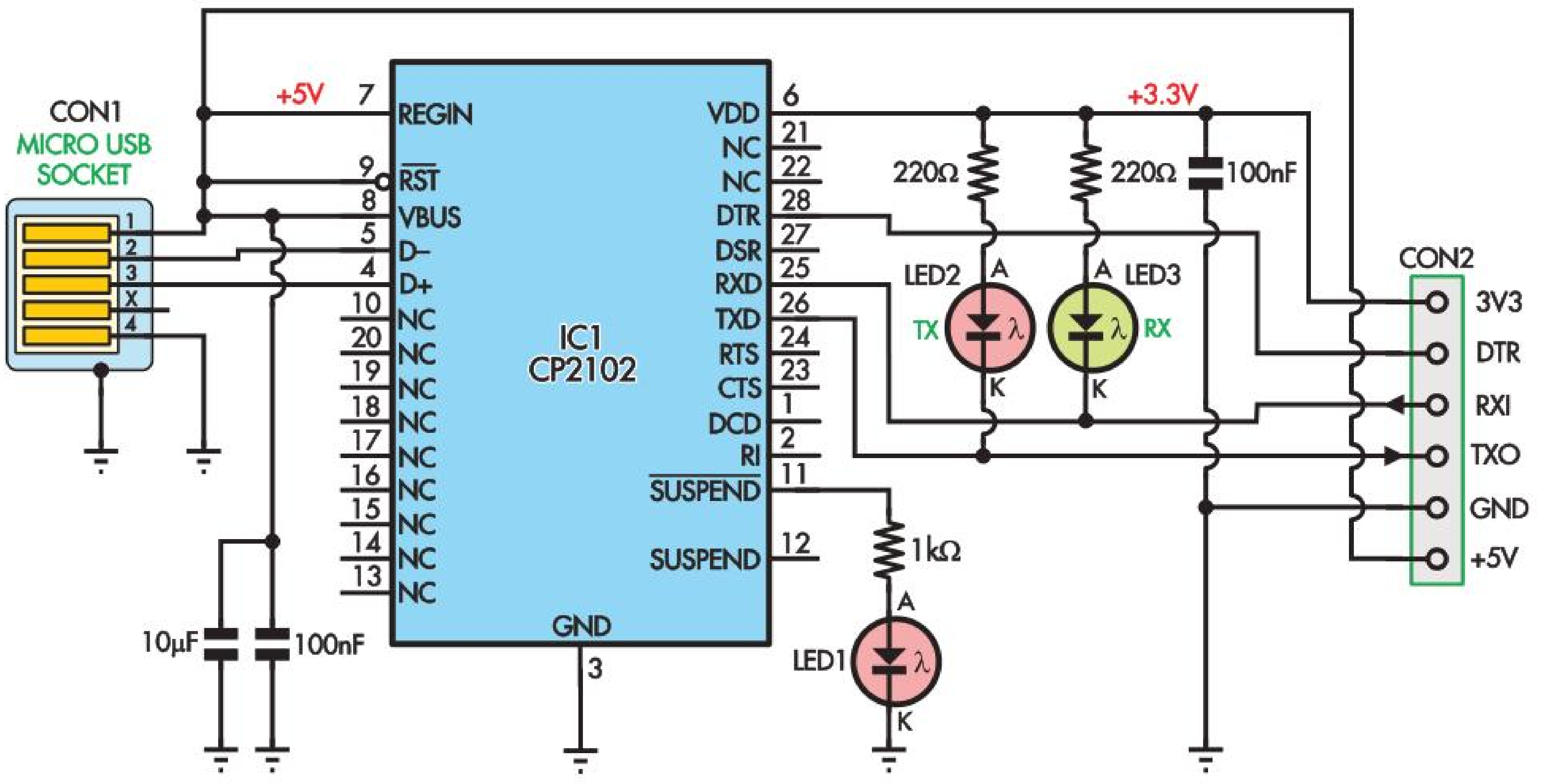


Fig.1: complete circuit diagram for the CP2102-based serial USB-UART bridge. The CP2102 can be powered directly from the USB V_{BUS} line and it contains a low drop-out voltage regulator to provide 3.3-3.45V (V_{DD}) from 4-5.25V (REGIN).

used to store the USB ID information: the vendor ID, the product ID, the serial number, the power descriptor, the release number and product description strings.

In addition, there are two RAM buffers, one 640-byte USB transmit buffer and one 576-byte USB receive buffer.

Since the CP2102 has a calibrated 48MHz oscillator, it needs no external crystal to operate at the USB 2.0 full-speed rate of 12Mbps. Finally, it contains its own low drop-out (LDO) voltage regulator, to give an output of 3.3-3.45V from an input (REGIN) within the range 4.0-5.25V. This means that it can be powered directly from the USB V_{BUS} line.

Circuit details

While this regulator can supply up to 100mA, the circuitry within the chip itself draws only a little over 26mA (maximum) even in normal operation, and only 100µA when suspended. This means it can supply up to 70mA or so for external circuitry needing a 3.3V supply.

In short, the CP2102 is a very impressive chip. Now turn your attention back to the module's circuit of Fig.1. There's a micro-USB socket at the left (CON1) to connect to a PC's USB port via a standard cable and also to power the module itself. So the V_{BUS} line from pin 1 of the socket connects to pins 7, 8 and 9 of the CP2102, with 10µF and 100nF bypass capacitors.

Note that the module does not provide connections to any of the CP2102 UART's handshaking lines, except for DTR ('data terminal ready'). However, this is unlikely to pose a problem for most applications nowadays, since even the DTR line is rarely used.

On the right-hand side there's a 6-way pin header (CON2) for the UART input, output and handshaking (DTR) connections, plus the ground, +5V and +3.3V power connections for use by external circuitry. There's also a 100nF bypass capacitor on the +3.3V line, plus three small indicator LEDs, each with its own series resistor for current limiting.

LED1 is driven from pin 11 of the CP2102, the $\overline{\text{SUSPEND}}$ output, so it only glows when the device is not suspended by the host PC, ie, when it's communicating with the PC normally via USB. On the other hand, LED2 and LED3 are connected between the +3.3V supply (pin 6) and pins 26 (TXD) and 25 (RXD) respectively, to indicate when data is being sent and received via the bridge.

LED1 draws a little over 1mA when it's operating, while LED2 and LED3 will each draw about 5mA. Thus the LEDs could draw up to 11mA from the 3.3V supply (with full duplex serial communications, allowing LED2 and LED3 to light simultaneously) and this should be taken into account when figuring out how much reserve current is available for external circuitry.

How to use it

Using the CP2102 based USB-UART bridge module is very straightforward. But before you can do so, you may need to install a virtual COM port (VCP) driver on your PC. This is the software which takes care of buffering data to and from the bridge and setting up the UART. In Windows, it makes the UART appear as if it were a legacy COM port.

You can get the right VCP driver from the Silicon Labs website: **www.silabs.com/products/interface/Pages/interface-software.aspx**

You can also download the latest version of the CP2102 data sheet from: **www.silabs.com/support/Pages/document-library.aspx**

When you go there you'll find they can provide VCP drivers for not only Windows 7-10, but also for Windows 2000/XP/Vista/Server 2003, WinCE, Mac OS 9 and X, Linux (3.x.x and 2.6.x) and Android. They can also provide drivers for direct 'USB-Xpress' interfacing to the PC, as an alternative to using the VCP approach.

Note that most modern operating systems, including Windows 10 and the latest versions of Mac OS X and Linux, should already have a suitable VCP driver installed. In this case, all you need to do is plug the bridge into a USB port and check that it has been recognised (eg, in Windows, check that a new COM port appears).

Once the driver is installed and working, you can set up your applications to communicate with the module via the new COM port. That includes setting the correct baud rate and other options.

Of course, your circuitry on the UART side of the module needs to be connected to the appropriate pins on header CON2. These will usually be just the RXI, TXO and GND pins, although you might also want to make use of one of the power supply pins as well.

If you aren't sure whether the bridge is working properly, the simplest way to test it is to wire up the RXI pin to the TXO pin. You can then open a terminal emulator, connect to that port

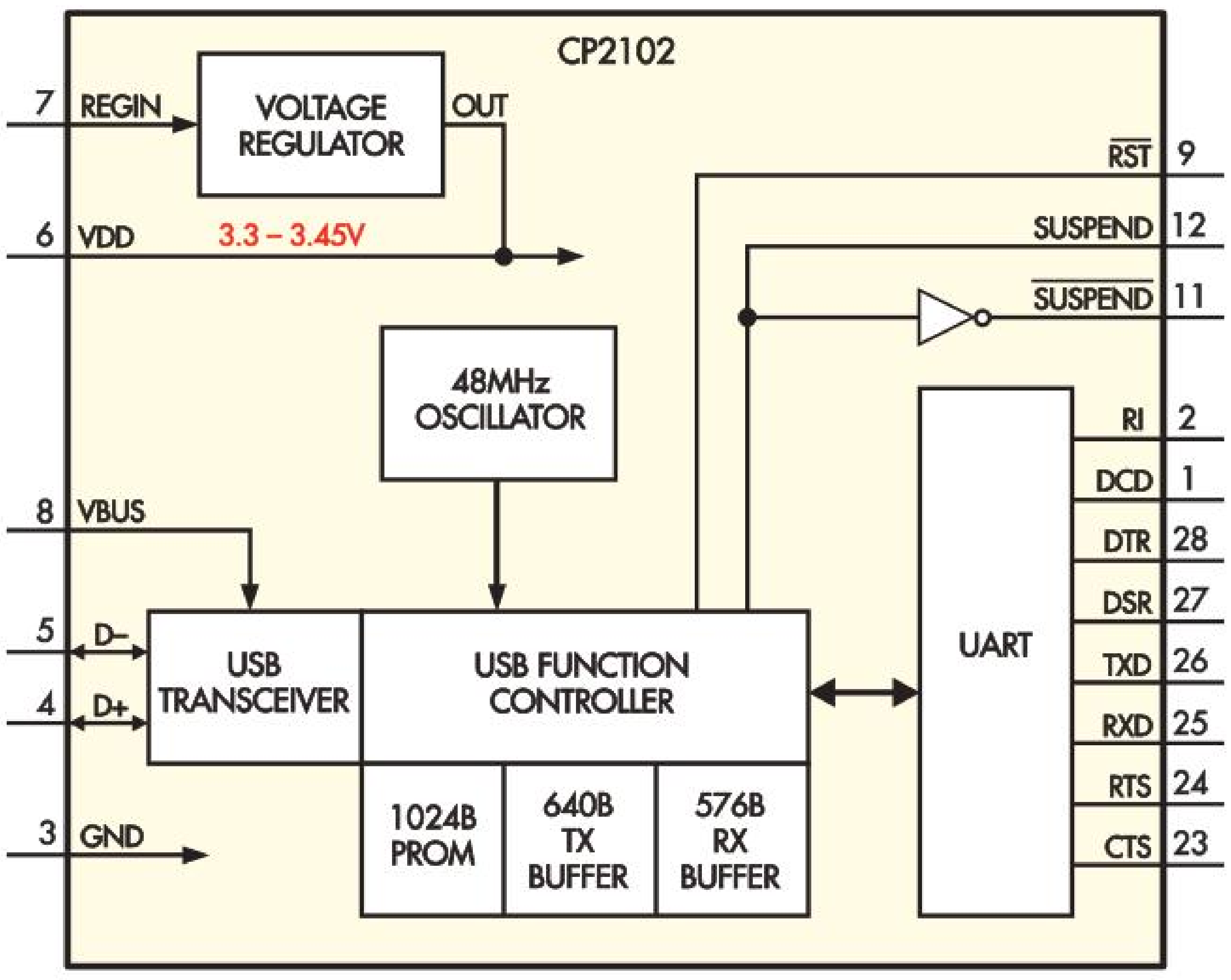


Fig.2: block diagram for the CP2102. This UART interface implements all RS-232 signals, including those for control and handshaking, although an external level shifter is required for full RS-232 compatibility.

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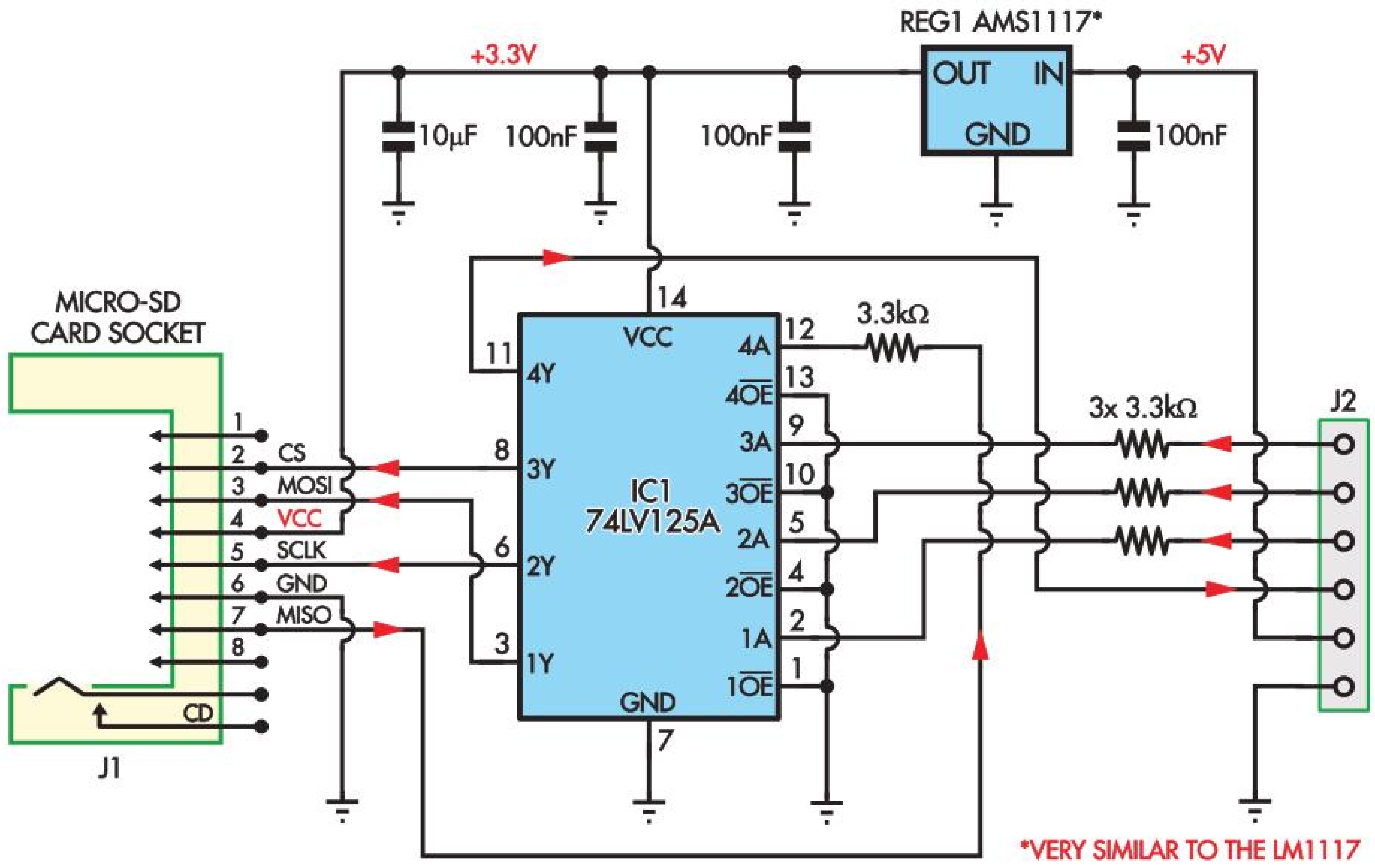


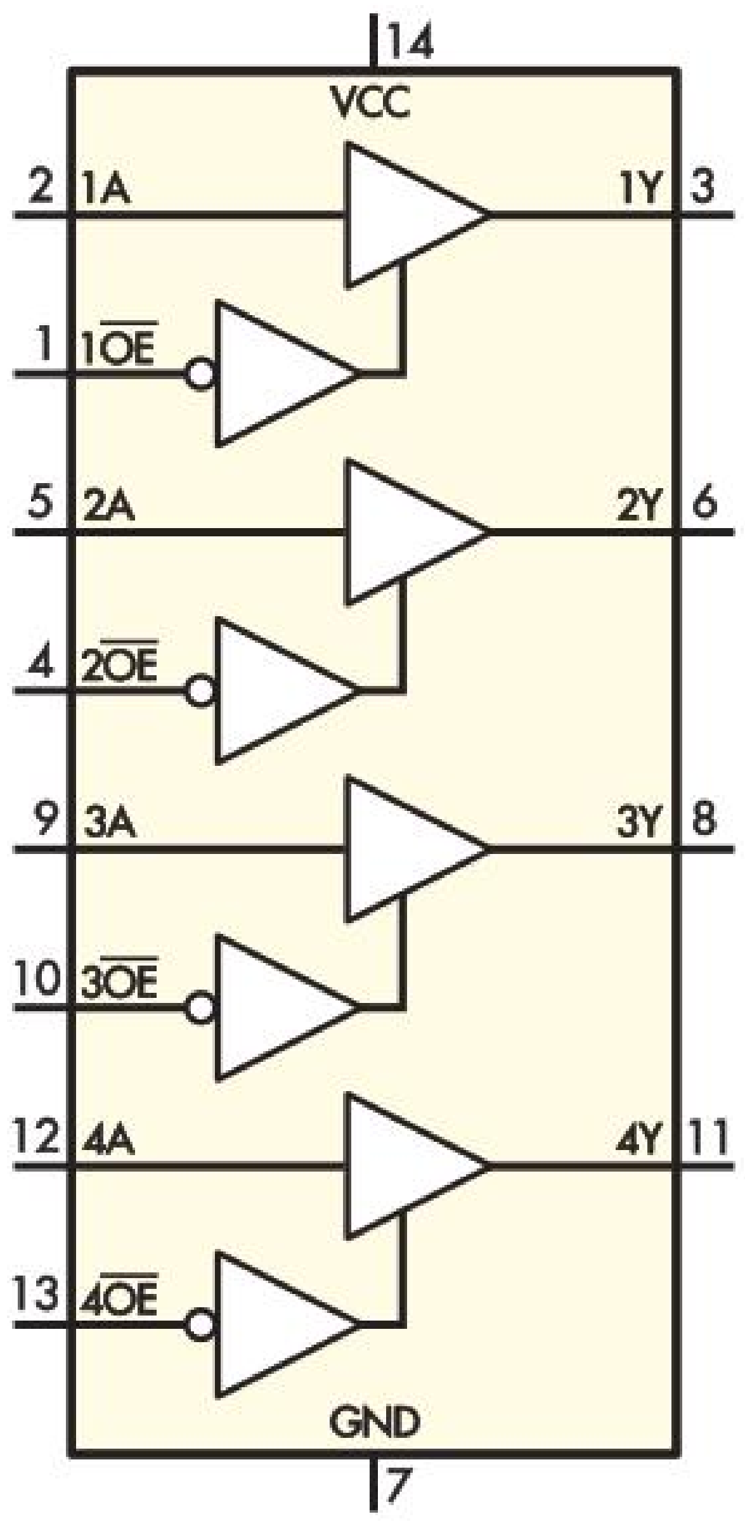


Fig.3: full circuit of the SPI/microSD adaptor module. REG1 reduces the 5V (V_{CC}) input supply from the host module to 3.3V, as required by microSD cards, while IC1 similarly reduces signal levels from the micro (which may run off 5V) to the 3.3V signal levels used by the SD card's I/Os.

Fig.4: internal block diagram of the SN74LV125A IC. When an $\overline{\text{OE}}$ input is pulled high, the corresponding output is disabled and has a high impedance.

and type on your keyboard. The typed characters should be sent back to you and appear in the terminal. If that works, but you still can't communicate with your target device, check that the connections to its TX/RX pins are not swapped and also that you have set the right baud rate.

microSD card interface

There are many different adaptors for accessing an SD memory card from a microcontroller or embedded module but they generally function in the same manner. The main differences are in terms of the card socket they provide and the chip(s) they use for interfacing.

The full circuit for this module is shown in Fig.3. Note that all SD cards can communicate via either serial peripheral interface (SPI) or a faster method, which consists of either a 4-bit parallel bus (older cards) or a high-speed differential interface (UHS-compatible cards). The SPI method is by far the simplest to implement with a microcontroller, unless it has a built-in SD card interface.

The other important thing to note is that all SD memory cards are intended to run from a 3.3V power supply and expect logic signals no higher than +3.3V. Some cards can only accept signals swinging over a smaller range, like 0-1.8V (UHS-I) or 0-0.4V (UHS-II).

Glossary

COM Port: PC communications port, normally sending and receiving data using the RS-232 serial protocol.

CS (Card/Chip Select): used in an SPI bus to indicate when the master wants to communicate with a slave (pulled low).

DTR (Data Terminal Ready): a 'flow control' signal which is used to indicate when the serial port is ready to receive data. Other, related flow-control signals include DSR (Data Set Ready), CTS (Clear To Send) and RTS (Ready To Send).

EEPROM (Electrically Eraseable, Programmable Read-Only Memory): non-volatile memory that can be erased and rewritten by applying a higher voltage than is used to read data back. EEPROM is normally more robust than Flash.

LDO (low drop-out [regulator]): a regulator which can maintain regulation with less than 2V between its input and output.

Micromite: a Microchip PIC32 programmed with the MMBasic interpreter.

MISO (master in, slave out): the serial data line used to transmit data from the selected slave to the master in an SPI bus.

MOSI (master out, slave in): the serial data line used to transmit data from the master to the selected slave in an SPI bus.

QFN (Quad Flat No-lead): a standard series of surface-mount integrated circuit packages. As the name suggests, it is attached to a PCB without through-holes via lands (pads) on the bottom and sides of the package (ie, without leads).

RS-232 or EIA-232: one of the most common standards for serial communications. Used by the serial ports on older PCs. Uses one wire for self-clocked data in each direction plus optionally, several flow control signals.

RX or RXD: serial data receive line. Normally connected to TX or TXD on the other device.

Serial Communication: the process of transferring data one bit at a time over a communication channel or bus.

SCK (Serial Clock): the shared clock line in an SPI bus, driven by the master, typically up to 20MHz.

SD (Secure Digital): a non-volatile portable storage device using Flash memory. Successor to MMC (MultiMedia Card).

SPI (Serial Peripherial Interface): a standard serial interface bus, commonly used between a microcontroller and peripherals such as SD cards. Unlike RS-232, SPI has a separate clock line – ie, three wires for bidirectional communications.

TTL (Transistor-Transistor Logic): refers to digital signals with a 5V or (later) 3.3V amplitude, as used in early digital circuits.

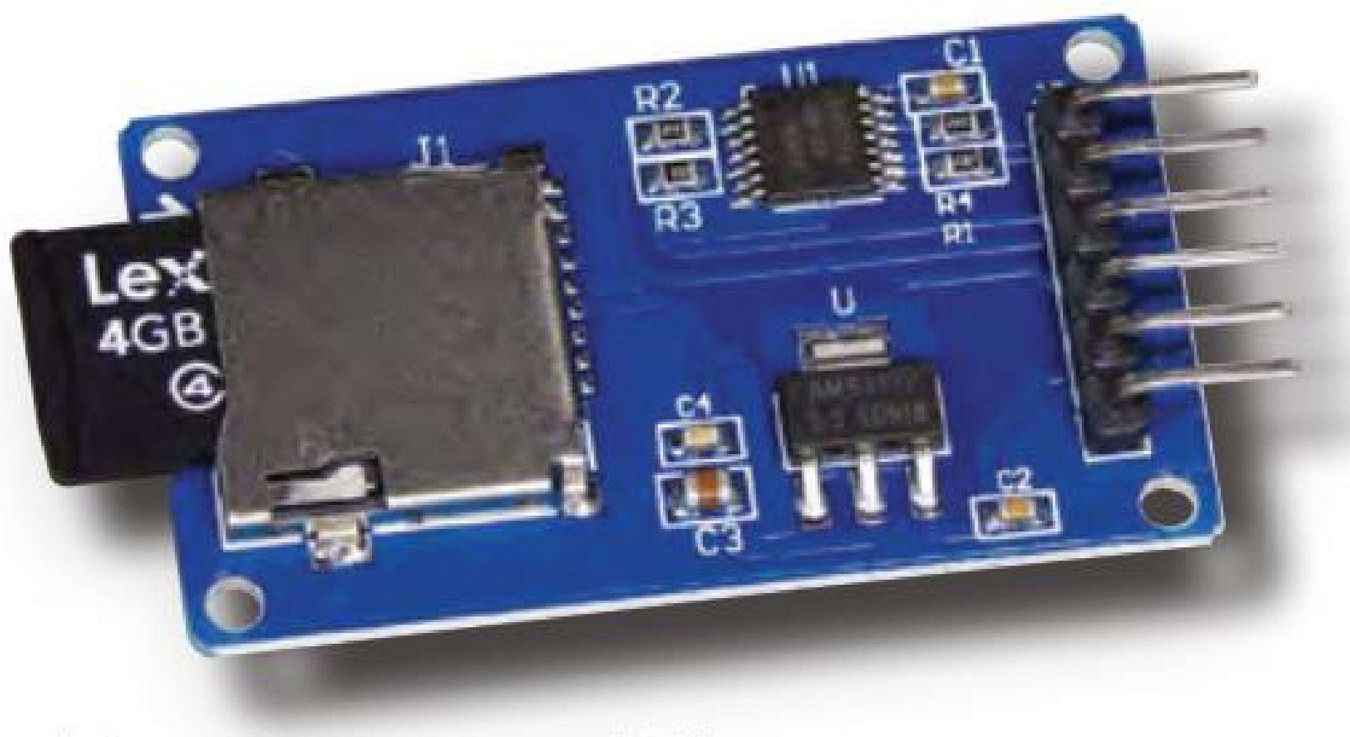
TX or TXD: data transmission line. Normally connected to RX or RXD on the other device.

UART (Universal Asynchronous Receiver/Transmitter): circuit to handle sending and receiving of serial data using one of several different serial protocols or variations thereof.

USB (Universal Serial Bus): high-speed serial bus with power (initially using four conductors) which replaced RS-232 and parallel ports for interfacing a PC to pluggable peripherals; from 1.5Mbps up to 5Gbps in latest version.

UHS (Ultra High Speed): transfer speed for the latest SD cards; up to 104MB/s for UHS-I, and 312MB/s for UHS-II.

VCP (Virtual COM Port): a device driver that emulates an RS-232 serial port over a different protocol such as USB.



This microSD module on a 43 × 24mm PCB is available from the SILICON CHIP online shop at: www.siliconchip.com.au/Shop/7/4019

Just because a chip has an SPI interface doesn't mean it can necessarily interface directly with an SD card. If the micro operates from a 5V supply, its SPI port(s) may well provide and expect logic high signals above +3.3V. This means that the adaptor is needed both to drop the supply voltage down to 3.3V (assuming a suitable rail is not already available elsewhere) and also to act as a logic-level translator for the SPI signals.

The module shown here incorporates LDO regulator REG1 to drop the +5V supply voltage from the micro (via J2) down to the +3.3V needed by both the microSD card at J1, and the single chip (IC1) on the module itself.

IC1 is an SN74LV125A tri-state buffer, to interface between the 5V logic levels (TTL) used on the micro side (via J2) and the low-voltage (0-3.3V) logic levels used on the SD card side (via J1). IC1 operates as a quad non-inverting buffer with tri-state outputs, ie, each output has its own $\overline{OE}$ (output enable low) input; see the internal block diagram of Fig.4. The $\overline{OE}$ inputs are not used, they are all tied to ground to enable the buffers permanently.

If you trace the signal paths through the circuit, you'll see that the three outgoing signal lines from the micro's SPI port at J2 (CS [card select], SCK [serial clock] and MOSI [data; master out, slave in]) each pass through a 3.3kΩ isolating resistor (to reduce ringing and provide some static electricity protection) and then through one of the buffers in IC1 to reach the corresponding pin on SD card socket J1.

For example, the 5V MOSI signal enters via J2, passes through its 3.3kΩ resistor and then goes to buffer input 1A (pin 2). The low-voltage logic version of this signal then emerges from the 1Y output (pin 3) and runs to the MOSI pin of J1, the microSD card socket.

The SCK and CS signals are processed via IC1 buffers 2 and 3 in the same way. The path followed by the MISO (data; master in, slave out) signal is similar, the only difference being that in this case the signal is travelling from the microSD card at J1 back to the micro at J2. Note though that the circuit does not level-shift this signal to 5V, so the micro will have to cope with a data input signal that only swings up to around 3.3V; most 5V micros are capable of this.

So the hardware side of the module is quite simple. Having said that, the SD card control protocol is quite complicated and so the software required to drive it is far from trivial.

Putting it to use

Since the module simply provides a transparent bridge linking the microSD card to the SPI port of your microcomputer, the software or firmware in the micro can exchange data with the card using the standard SPI commands. So with an Arduino, you can use commands like:

```
SPI.beginTransaction(SPISettings());
receivedVal = SPI.transfer(val);
SPI.end();
```

There's also an Arduino code library built into recent versions of the Arduino IDE, designed especially for reading from and writing to SD cards. It offers commands like begin(), mkdir(), open(), remove(), rmdir(), available(), close(), write() and read().

With a Micromite it's also fairly straightforward, using commands such as:

```
SPI OPEN speed, mode, bits
received_data = SPI(data_to_send)
SPI CLOSE
```

However, the Micromite Plus has built-in library commands specifically intended for reading and writing to SD cards; see the recent articles on Micromite programming.

Useful links

- Information on using standard SPI commands with an Arduino, including some short examples, can be found at: **www.arduino.cc/en/Reference/SD**
- Details on using SPI communications with a Micromite begin on page 92 of the Micromite manual: **http://geoffg.net/Downloads/Micromite/Micromite%20Manual.pdf**
- An article on the SPI bus is available at: **http://en.wikipedia.org/wiki/Serial_Peripheral_Interface_Bus**
- Wikipedia also has a very informative article on the many kinds of SD cards, at: **http://en.wikipedia.org/wiki/Secure_Digital**

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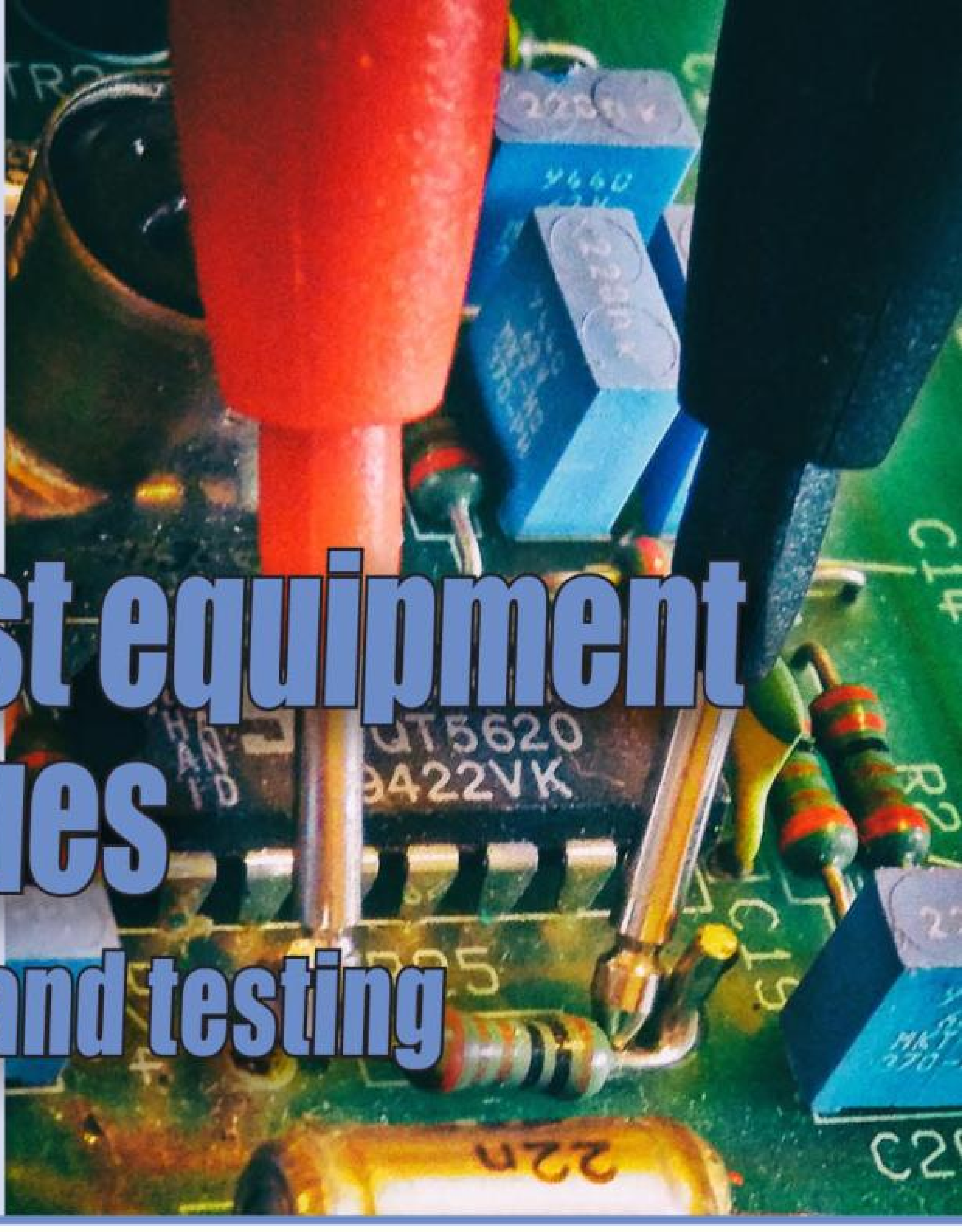
Teach-In 2018

Get testing! – electronic test equipment and measurement techniques

Part 6: Audio frequency measurement and testing



by Mike Tooley

Welcome to *Teach-In 2018: Get testing! – electronic test equipment and measurement techniques*. This *Teach-In* series will provide you with a broad-based introduction to choosing and using a wide range of test gear, how to get the best out of each item and the pitfalls to avoid. We'll provide hints and tips on using, and – just as importantly – interpreting the results that you get. We will be dealing with familiar test gear as well as equipment designed for more specialised applications.

Our previous *Teach-In* series have dealt with specific aspects of electronics, such as PICs (*Teach-In 5*), Analogue Circuit Design (*Teach-In 6*) or popular low-cost microcontrollers (*Teach-In 7* and *8*). The current series is rather different because it has been designed to have the broadest possible appeal and is applicable to all branches of electronics. It crosses the boundaries of analogue and digital electronics with applications that span the full range of electronics – from a single-stage transistor amplifier to the most sophisticated microcontroller system. There really is something for everyone in this series!

Each part includes a simple but useful practical *Test gear project* that will build into a handy gadget that will either extend the features, ranges and usability of an existing item of test equipment or that will serve as a stand-alone instrument. We've kept the cost of these projects as low as possible and most of them can be built for less than £10 (including components, enclosure and circuit board).

This month

In this, the sixth part of *Teach-In 2018*, *In theory* looks at the specifications used for audio frequency (AF) equipment, including power amplifiers, preamplifiers, filters and tone controls. *Gearing up* introduces AF test equipment and techniques, while *Get it right!* helps you avoid test and measurement pitfalls and provides useful hints and tips that will help you to improve the accuracy and relevance of your measurements. Finally, our sixth *Test gear project* is a low-distortion 1kHz signal source that can be used to carry out a variety of useful tests and measurements.

In theory: AC measurement

Several different parameters are used in performance specifications for audio equipment such as power amplifiers, preamplifiers, filters and tone controls. Many are obvious, but a few need a little explanation. Those that readers will doubtless already be familiar with include:

- Voltage gain (expressed as a voltage ratio or in dB)
- Input and output impedance
- Frequency response (or bandwidth limits)
- Output power.

Several other parameters may be less obvious – they include:

- Phase/frequency response
- Damping factor (DF)
- Transient response (rise/fall time and slew rate)
- Cross-talk
- Distortion (HD, THD and IMD)
- Hum and noise
- Signal-to-noise ratio (SNR).

Not every parameter is important in a particular application, so before we look at how these specifications are measured it is worth explaining what each means and how it impacts on the performance of a particular item of audio equipment. To put these parameters into context, Table 6.1 lists the specifications

Table 6.1 Specification for a high-performance audio power amplifier

Parameter	Quoted specification
Number of channels	Two (note 1)
Rated power output	75W into 8Ω at less than 0.03% THD (notes 2 and 3)
Voltage gain	28dB
Input sensitivity for rated output	1V
Frequency response	20Hz to 20kHz at -1dB and 5Hz to 100kHz at -3dB
Phase response	+5°/–15° over the frequency range 20Hz to 20kHz
Signal-to-noise ratio	Greater than 120dB over the range 20Hz to 20kHz
Total harmonic distortion (THD)	Less than 0.03% (note 4)
Intermodulation distortion (IMD)	Less than 0.03% (note 4)
Rated load impedance	4Ω to 16Ω
Damping factor	Greater than 400
Slew rate	Better than 50V/µs
Input impedance	28kΩ
DC output offset	Less than ±5mV

Notes

1. Left (L) and right (R) channels of a stereophonic system
2. Per channel with both channels driven
3. Full-bandwidth power (FTC) is 60W
4. At all power levels up to, and including the rated output power.

Fig.6.1. Some prototype amplifiers, filters and tone controls ready for testing

of a high-performance audio power amplifier.

Voltage gain

The voltage gain provided by an amplifier is simply the ratio of output voltage to input voltage (where input and output voltages are both specified in the same units – eg, both RMS or both peak-peak values).

Input sensitivity

The input sensitivity of an amplifier is the input voltage needed to produce the amplifier's rated output.

Input and output impedance

The input impedance of an amplifier is the impedance 'seen' looking into the amplifier's input. Similarly, the output impedance of an amplifier is the impedance 'seen' looking back into the amplifier's output. This means that the output impedance of an amplifier is the internal impedance of the amplifier at its output. On page 38 we explain this concept in more detail by using an equivalent circuit.

Frequency response

The frequency response of an amplifier is usually expressed in terms of its lower and upper cut-off frequencies. 'Cut-off' is a confusing term since it might imply that there is no amplification below and above the lower and upper cut-off frequencies. This is not the case, and the term simply refers to a given reduction in output voltage or power. In most cases, the cut-off frequency is taken to mean the frequency at which the output voltage has fallen to 70.7% of its mid-band value (this is the point at which the output *power* falls to exactly half of the mid-band value).

Output power

The output power produced by an amplifier is the maximum power that it can deliver under linear conditions (ie, with a sinusoidal input and output). When this is the case we can express the power delivered to a load (of given impedance or resistance) in RMS watts. Since distortion increases with output power this parameter is often specified at a specific level of distortion (eg, 10W for 0.1% distortion or 15W for 1% distortion).

To overcome the ambiguities associated with the specification of audio output power, several standards have been introduced. The notion of 'continuous average power' was first introduced in the US and several other countries as a means of dispelling some of the myths associated with audio power measurement and its specification. The term defines the output of an amplifier when delivered on a sustained basis and the Federal Trade Commission (FTC) standard, originally published in 1974, requires that, before measuring the output power, an amplifier should be 'preconditioned'. The FTC states that this process should involve operating the amplifier with all channels simultaneously driven at one third of rated output power for a period of one hour. The FTC standard also specifies the temperature (25°C) at which the measurement should be carried out and further states that it should be carried out at 'all frequencies within the power band' and 'without exceeding the rated maximum percentage of total harmonic distortion' (THD). Importantly, the FTC requires that the rated power output should be sustainable for a period of not less than five minutes.

The most recent international standard for specifying amplifier performance is defined in IEC (BS EN) 60268 Part 3, *Sound system equipment: Amplifiers*. Published in 2013, this standard applies not only to conventional analogue amplifiers but also to the analogue parts of analogue/digital amplifiers that form part of a sound system for professional or household applications. The standard specifies the characteristics that should be included in specifications of amplifiers and the corresponding methods of measurement. In particular, it defines methods for measuring the short-term, long-term and temperature-limited output power of an amplifier.

Phase response

The output signal of an amplifier may not rise and fall in sympathy with its input, and there will often be an amount of angular shift (expressed in degrees) between the two signals. For example, if the output signal from an amplifier is inverted when compared with its input there will be 180° of phase shift between the input and output signals. The input signal is usually taken to be the reference signal and the output may shift forwards (leading) or backwards (lagging) over a range of frequencies.

Damping factor

The output impedance of a power amplifier is usually very small (a fraction of an ohm) but is rarely specified and is difficult to measure. Instead, damping factor (DF) is used as a measure of the output impedance relative to that of the load (usually a much higher value). DF is simply the ratio of the load impedance (often 4Ω, 8Ω or 15Ω) to the amplifier's internal impedance (usually less than a tenth of an ohm). In matched systems (such as 600Ω line amplifiers) the input and output resistances will both be the same, yielding a unity DF.

Transient response

Transient response is important in high-quality and wideband amplifiers, and it can be expressed in several different ways including rise/fall time and slew rate. Rise time is the time taken for the output voltage to swing from its most negative value to its most positive value, whereas fall time is the time taken for the output voltage to swing from its most positive value to its least positive value. Rise time is conventionally measured between the 10% and 90% points of the leading edge of a positive-going transient, while fall time is measured between the 10% and 90% points of a falling-edge transient. Both times should be very small (ideally no more than a few tens or hundreds of nanoseconds). An alternative measure is based on the rate at which a perfect (ie, zero time) transient voltage rises or falls at the output of an amplifier. In effect, the specified parameter is the rate of change of voltage with time.

Cross-talk

When more than one channel is present within an amplifier there may be some leakage of signal from one channel into the other. This can arise from various causes, including inadequate screening and poor power supply design. Cross talk is usually measured in decibels (dB) and is the ratio of signal from one channel appearing in the other. Cross-talk figures are usually high, with values of 90 to 120dB being typical.

Distortion

The two most common forms of non-linear distortion are harmonic distortion (HD) and intermodulation distortion (IMD). HD is often specified in terms of the *total* amount of harmonic distortion present (THD). In both of these forms of distortion extra frequency components are added to a signal due to non-linearity of an amplifier's transfer characteristic. If an amplifier had a perfectly linear transfer characteristic there would be no HD and no IMD.

Distortion is usually expressed in decibels (dB) or as a percentage of the rated output. Even-order harmonic distortion is generally caused by an asymmetrical transfer characteristic whereas odd-order harmonic distortion is caused by a symmetrical non-linearity. IMD arises from different signal frequency components mixing together to produce new frequency components.

Hum and noise

Hum and noise within an amplifier can also be added to a signal. These are also unwanted components present at the output that are not present at the input. Hum is simply the appearance of a signal at mains supply frequency (or twice the mains supply frequency). Hum can be carried on supply voltage rails, where it appears as a small AC signal

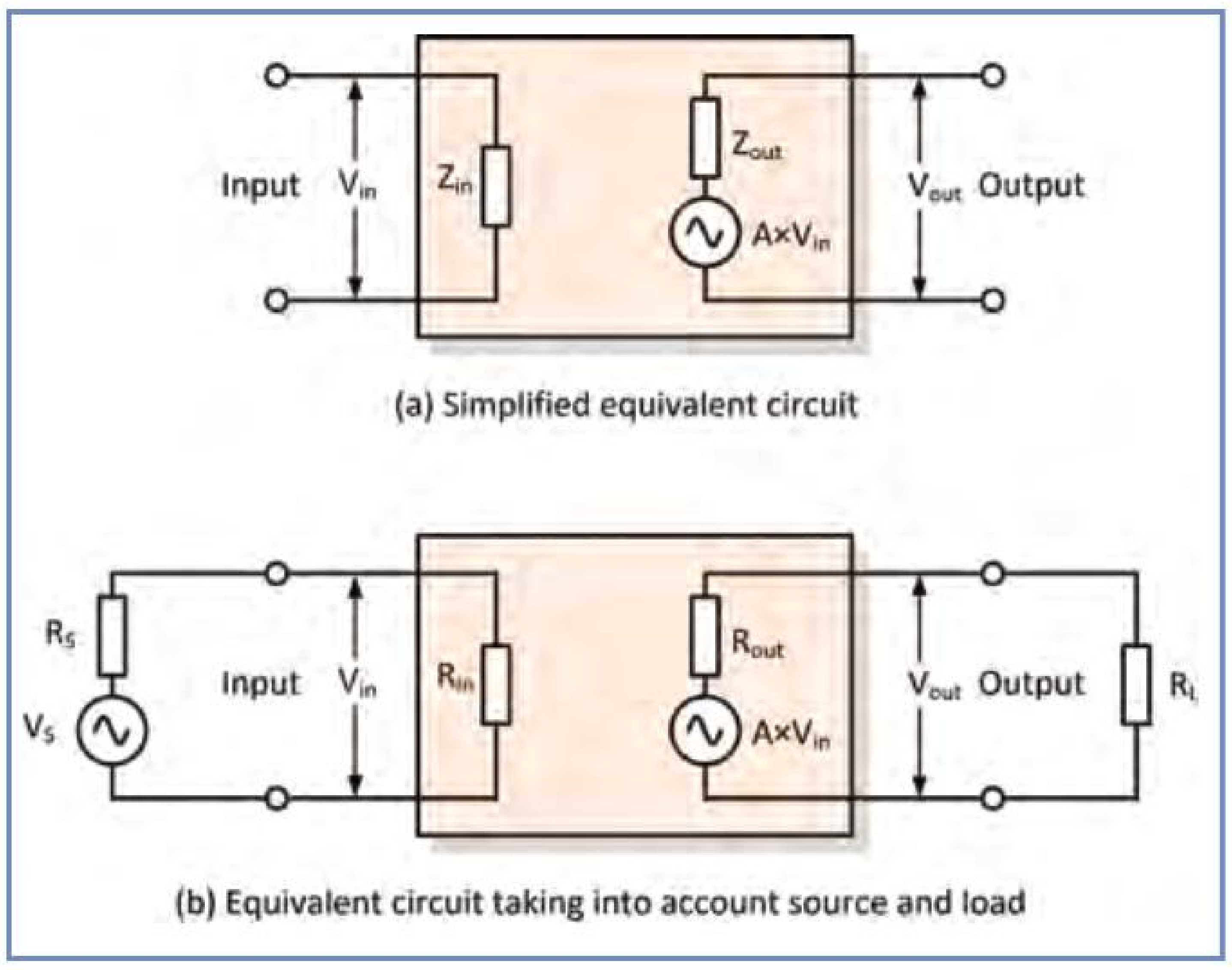

Input
V_{in}
Z_{in}
Z_{out}
A×V_{in}
V_{out}
Output
(a) Simplified equivalent circuit
R_S
V_S
Input
V_{in}
R_{in}
R_{out}
A×V_{in}
V_{out}
Output
R_L
(b) Equivalent circuit taking into account source and load

Fig.6.2. Equivalent circuits of an amplifier

superimposed on the DC supply. It can also find its way into an amplifier when stray magnetic fields (such as those that surround power transformers) induce current into nearby wiring. Hum can be reduced, if not completely eliminated, by good power supply design, using screened signal cables, and by adequate screening and grounding of chassis and chassis-mounted components.

Noise is a random fluctuation superimposed on a wanted signal. Unfortunately, all electronic components produce noise – but some produce more noise than others. The amount of noise that they produced depends not only on the type, construction and material used in the component, but also on the electrical conditions under which it is operated (ie, current and voltage) as well as, very significantly, the temperature. Noise is a particular problem within high-gain amplifiers where noise generated in the first stage receives the full benefit of the gain provided by the subsequent stages.

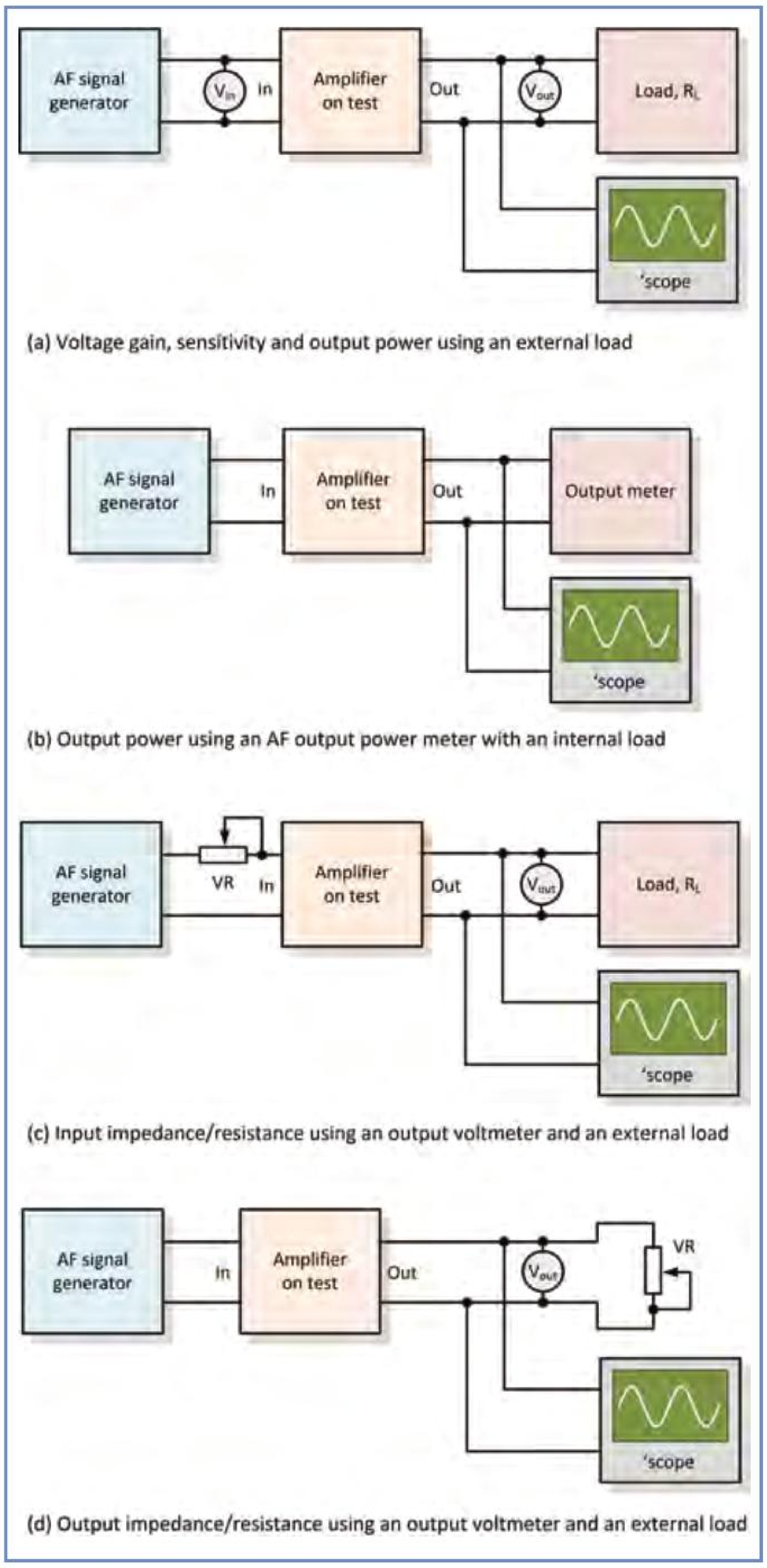

AF signal generator
V_{in}
In
Amplifier on test
Out
V_{out}
Load, R_L
'scope
(a) Voltage gain, sensitivity and output power using an external load
AF signal generator
In
Amplifier on test
Out
Output meter
'scope
(b) Output power using an AF output power meter with an internal load
AF signal generator
VR
In
Amplifier on test
Out
V_{out}
Load, R_L
'scope
(c) Input impedance/resistance using an output voltmeter and an external load
AF signal generator
In
Amplifier on test
Out
V_{out}
VR
'scope
(d) Output impedance/resistance using an output voltmeter and an external load

Fig.6.3. Arrangements for measuring basic amplifier specifications

Signal-to-noise ratio

Signal-to-noise ratio (SNR) is the ratio of wanted signal to the amount of noise present and is usually expressed in decibels (dB). Since distortion will invariably also be present and can be difficult to separate during measurement, a more practical measure is often used. SINAD, or 'signal in noise advantage', is a measure of the signal quality that takes into account the presence of distortion. SINAD is calculated from the ratio of total output signal power (ie, signal power plus noise power plus the power arising from distortion components) to the noise-plus-distortion power.

As might be expected, SINAD performance is only slightly worse than the SNR performance but the SINAD figure is usually a more reliable measure of performance, particularly when a significant amount of noise and distortion is present. Note that, since the response of the human ear favours the middle frequency range (from about 500Hz to around 5kHz), noise is often specified A-weighted in relation to its effect within this band.

Equivalent circuit of an amplifier

Before making meaningful measurements on an amplifier it is useful to have an idea of what goes on 'inside the box' and what happens when the 'box' is connected to a source and load. For this, we can make use of an equivalent circuit, like that shown in Fig.6.2(a). This simplifies the internal circuit to just three components and ignores the effect of changes at the output affecting the input. The three components present are:

- Input impedance, Z_{in}
- Output impedance, Z_{out}
- Voltage source, $A_V \times V_{in}$

When the amplifier is driven from a source and connected to a load, the equivalent circuit becomes a little more complicated and is shown in Fig.6.2(b). It assumes that in the mid-band frequency range the reactive components are negligible, so we've further simplified the circuit by replacing the impedances with their corresponding resistances. Thus, the voltage at the input of the amplifier will be:

$$V_{in} = V_{in} \times \frac{R_{in}}{R_{in} + R_S}$$

While at the output of the amplifier it will be:

$$V_{out} = (A_V \times V_{in}) \times \frac{R_L}{R_L + R_{out}}$$

If $R_{in} >> R_S$ and $R_{out} << R_L$ then the overall voltage gain (G) will be given by:

$$G = \frac{V_{out}}{V_{in}} = \frac{A_V \times V_{in}}{V_{in}} = A_V$$

Measuring amplifier specifications

Measuring voltage gain, sensitivity and frequency response

To measure the voltage gain, sensitivity and frequency response of an amplifier, a sinewave signal at a frequency in the centre of the amplifier's mid-band range (usually 1kHz) is applied to the input, and the amplifier is adjusted for an undistorted output by observing the output waveform using an oscilloscope. The amplifier should be connected to a suitably resistive load (not a loudspeaker) as shown in Fig.6.3(a). RMS-reading AC voltmeters should be used to measure the input and output voltages from which the voltage gain can be calculated (see earlier). Similarly, the sensitivity can be measured by adjusting the amplifier for its rated output power and measuring the corresponding input voltage (any gain or volume control fitted to the amplifier must usually be set to maximum before making this measurement).

The arrangement shown in Fig.6.3(a) can also be used to measure an amplifier's frequency response. The output frequency of the AF signal generator is adjusted over the amplifier's full working range and, at each end of this range the frequency at which the output voltage falls to 70.7% of its mid-band value is located. These are the two 'cut-off' frequencies. If the amplifier's response is

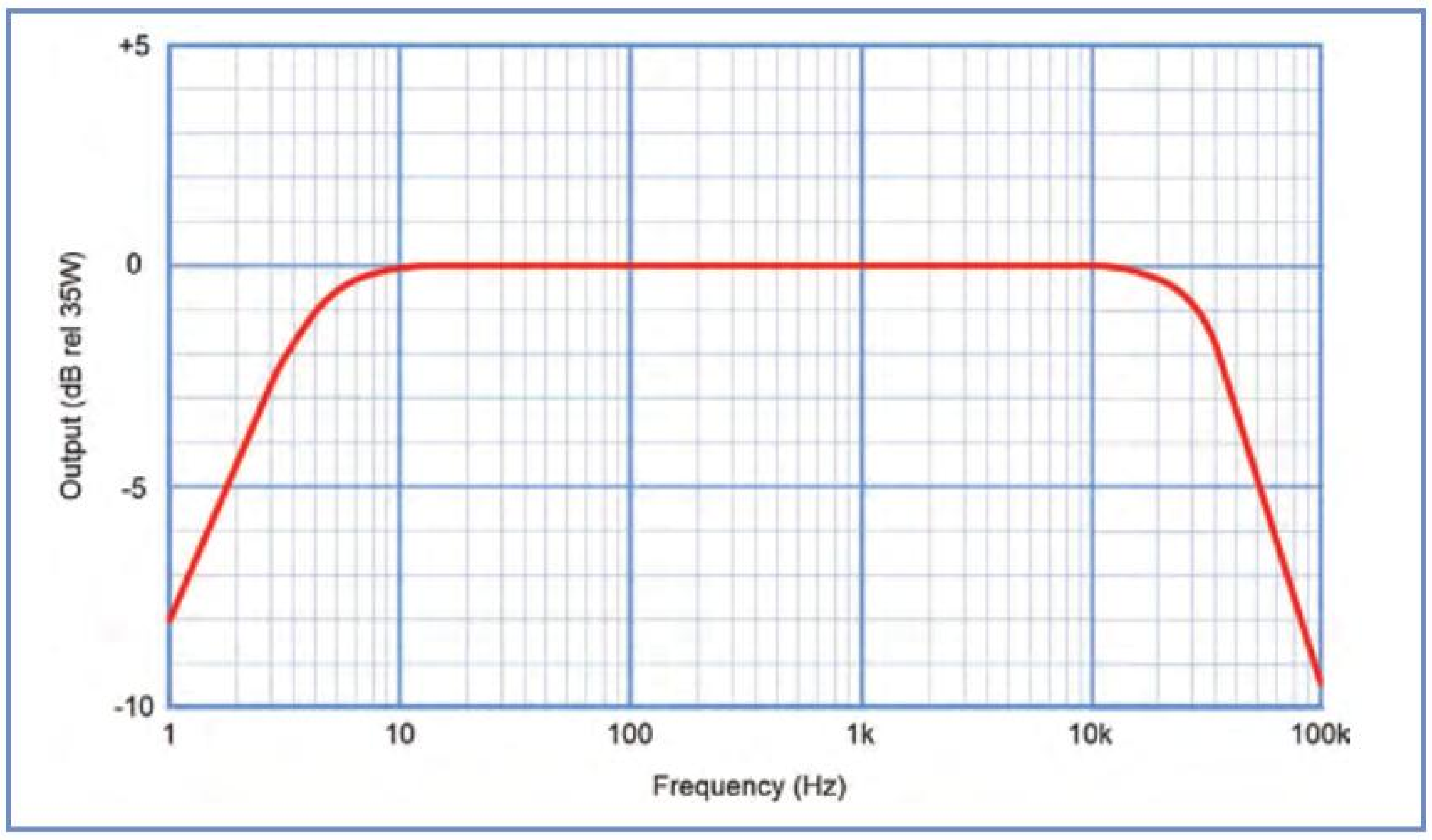


Fig.6.4. Measured frequency response of a Technics SU-Z22 power amplifier

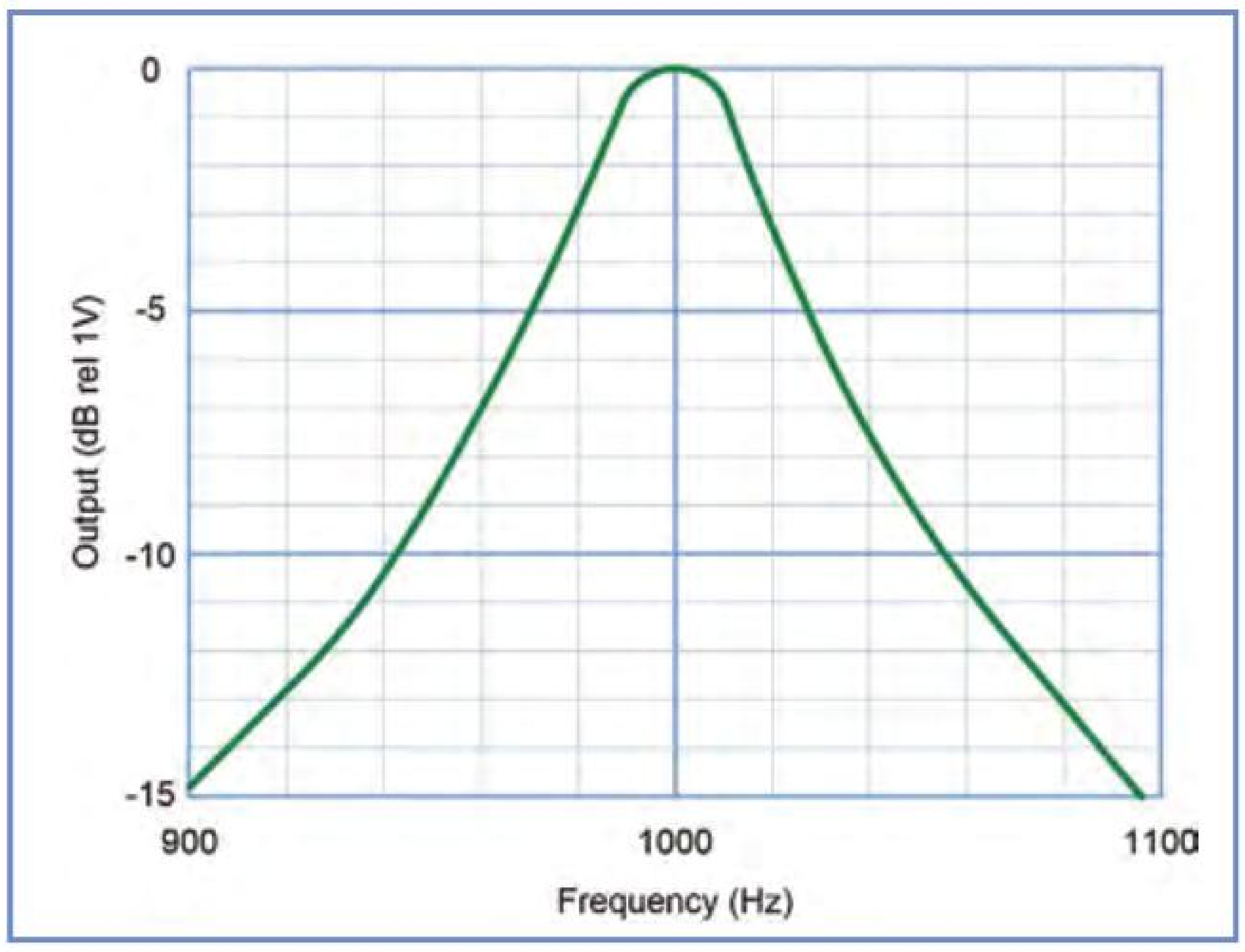


Fig.6.5. Measured frequency response of a 1kHz filter (ref: 0dB at 1kHz)

not substantially 'flat' then it is advisable to take a series of readings and plot the frequency response as a graph. Some typical frequency response graphs are shown in Fig.6.4 and Fig.6.5.

Output power

Output power should be measured under sinusoidal conditions and should be the power that the amplifier can deliver on a continuous basis. Since values of voltage and current measured are both expressed in terms of 'root mean square' quantities, this power is sometimes referred to as 'RMS power'. These values can be read from a voltmeter or ammeter calibrated for sinewave operation (as is invariably the case with instruments that are designed to make conventional AC power line measurements). The term 'RMS power' is, however, somewhat misleading since the power that is indicated by this measurement is actually the *average* power over a cycle of the wave. When this power is dissipated in a resistor it appears as heat.

To put this into context, let's assume that you have a load of known resistance (not a loudspeaker) connected to an amplifier, and that the amplifier is supplied with a sinewave signal source within the middle of the audio band (usually 1kHz). We also need to ensure that the output of the amplifier (the voltage or current that we are going to measure) is not distorted (ie, that it is truly sinusoidal). In order to do this we would need to observe the output waveform, checking with an oscilloscope that it has not become clipped or otherwise distorted. If we now measure the RMS AC voltage we can determine the output power (P_{out}) from:

$$P_{\text{out}} = I_{\text{out}} \times V_{\text{out}} = \left(\frac{V_{\text{out}}}{R_{\text{L}}}\right) \times V_{\text{out}} = \frac{V^2_{\text{out}}}{R_{\text{L}}}$$

If, for example, we had used an 8Ω load and measured an undistorted 10V RMS developed across it we would be able to determine the power output from:

$$P_{\text{out}} = \frac{V^2_{\text{out}}}{R_{\text{L}}} = \frac{10^2}{8} = \frac{100}{8} = 12.5\text{W}$$

If we had used an oscilloscope (instead of an AC meter) to measure the output voltage we would probably find it much easier to measure the peak-to-peak value of the waveform. We can then convert this value, $V_{\text{out(pk-pk)}}$, to an RMS value and use that value in our calculation, as follows:

$$P_{out} = \frac{V^2_{out}}{R_L} = \frac{\left(\frac{V_{out(pk-pk)}}{2\sqrt{2}}\right)^2}{R_L} = \frac{\left(V_{out(pk-pk)}\right)^2}{8R_L}$$

Putting this into context with some typical figures, let's assume that the oscilloscope indicates a peak-peak voltage of 20V with a load having a resistance of 4Ω. The output power would be calculated from:

$$P_{out} = \frac{\left(V_{out(pk-pk)}\right)^2}{8R_L} = \frac{20^2}{8 \times 4} = \frac{400}{32} = 12.5\text{W}$$

Measuring output power

The arrangement for measuring output power is similar to that used for voltage gain and sensitivity measurement, as shown in Fig.6.3(b). If an AF output power meter is available then this can usually provide the load required, but if such an instrument is unavailable a separate external load will be required. The test load should not only be purely resistive, but it should also be adequately rated in terms of power dissipation. For most practical purposes this means that one or more high-power resistors will be required, and using a suitable combination of series and/or parallel components it is possible to provide several different load resistances.

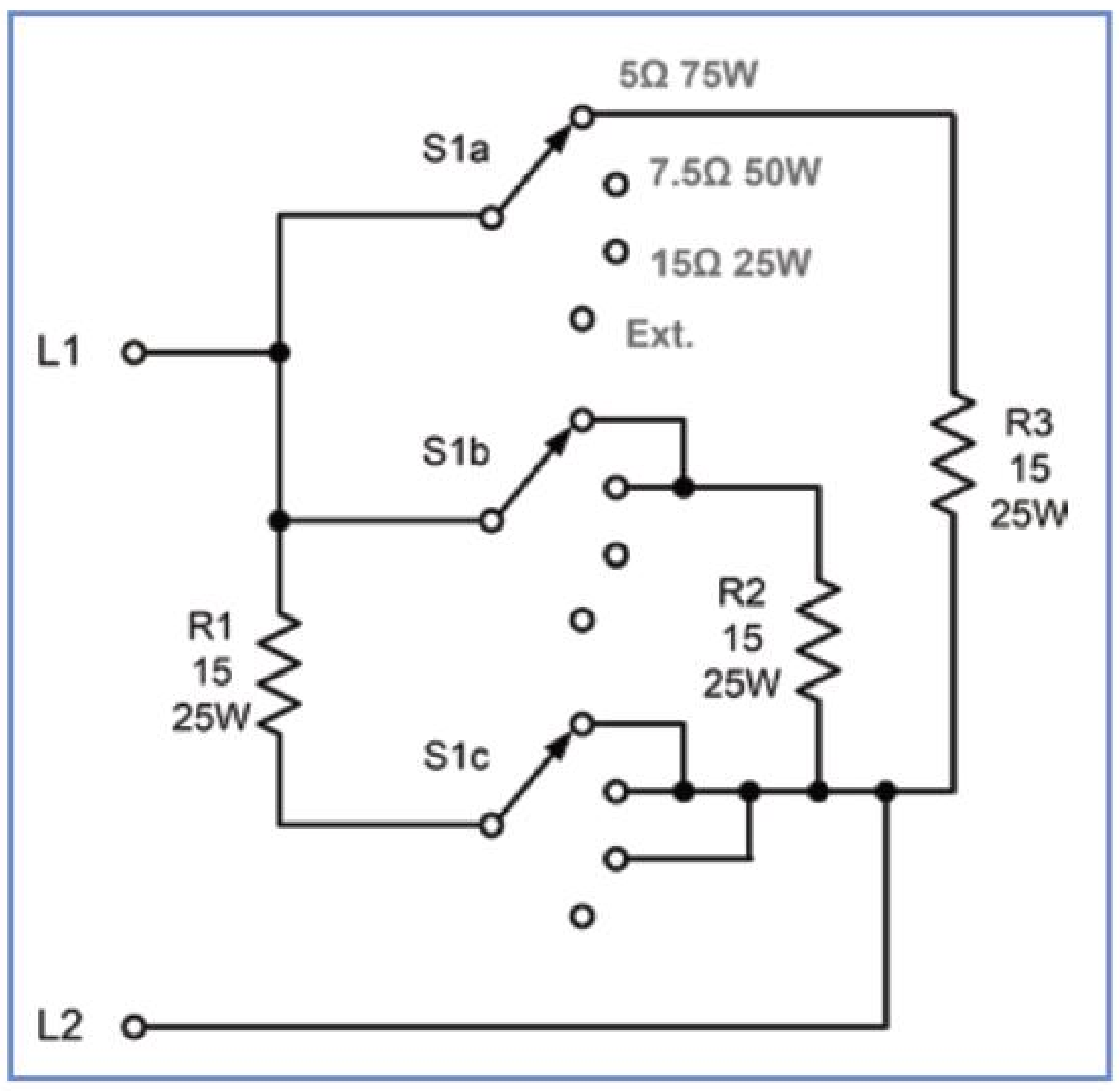


Fig.6.6. Circuit of the output load

The circuit of a simple but effective load is shown in Fig.6.6. The load can be switched to provide resistances of 5Ω, 7.5Ω and 15Ω at maximum power ratings of 75W, 50W and 25W respectively. To realise these values we used three high-power aluminium-clad wire-wound 15Ω resistors. Each resistor should be rated at 25W when mounted and used according to the manufacturers recommendations (note that the manufacturer suggests de-rating by 50% when the resistors are not used with a heat-dissipating surface (ie, a heat sink) or when they are used at a high ambient temperatures. The resistors are Arcol part number 'HS25 15RJ' (or similar) and they are available from several electronic component suppliers including Rapid Electronics (stock code 62-8430). Note that the switched 5Ω, 7.5Ω or 15Ω resistances offered by our output load should prove satisfactory with amplifiers rated for 4Ω, 8Ω or 15Ω loads.

Measuring input impedance/resistance

An arrangement for measuring input resistance is shown in Fig.6.3(c). A variable resistor or decade resistance box – 'VR' – is inserted at the input. This resistor is initially set to zero and the amplifier is adjusted for normal operation (without clipping or distortion evident from the waveform displayed on the oscilloscope). The output voltage is then noted, after which VR is adjusted until the output falls to exactly half this value. At this point, the value of resistance is measured (or read from the decade resistance box). The input

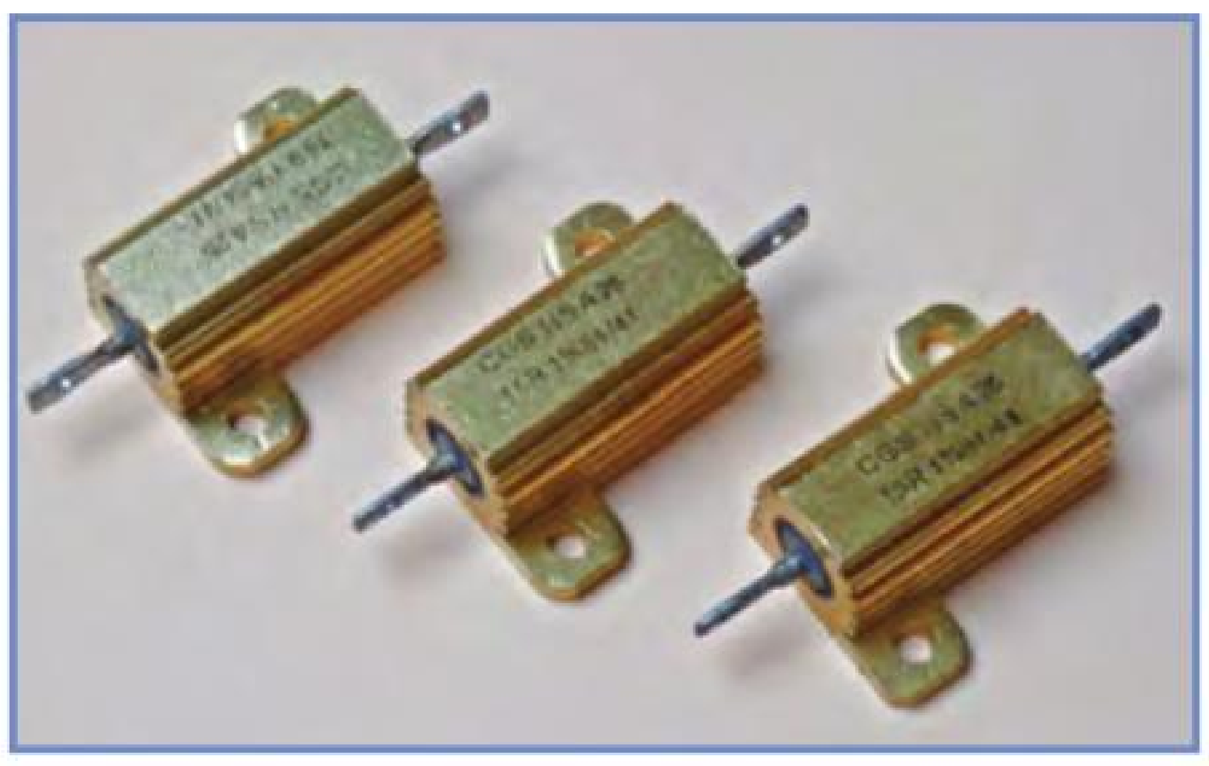


Fig.6.7. The three 15Ω 25W wire-wound resistors used in the switched output load

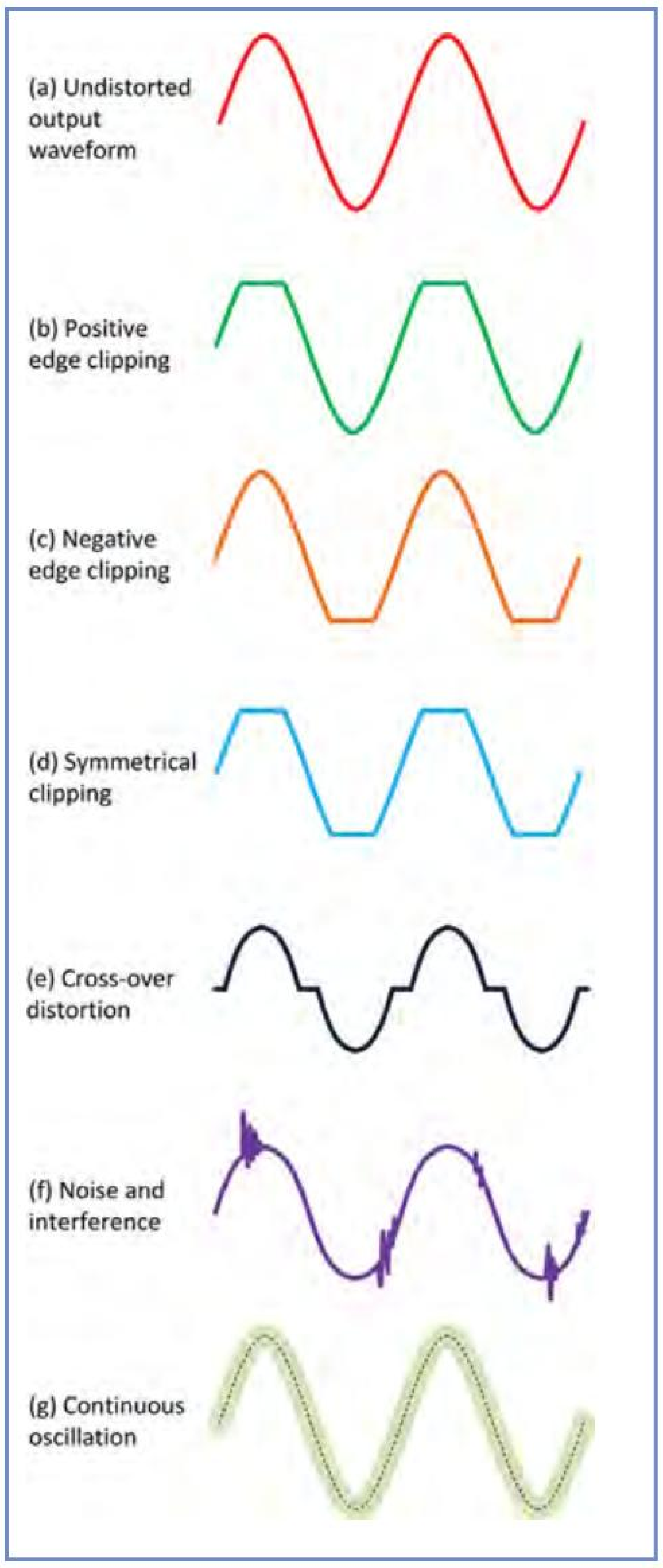

(a) Undistorted output waveform
(b) Positive edge clipping
(c) Negative edge clipping
(d) Symmetrical clipping
(e) Cross-over distortion
(f) Noise and interference
(g) Continuous oscillation

Fig.6.8. Using a sinewave signal to test for various distortion conditions

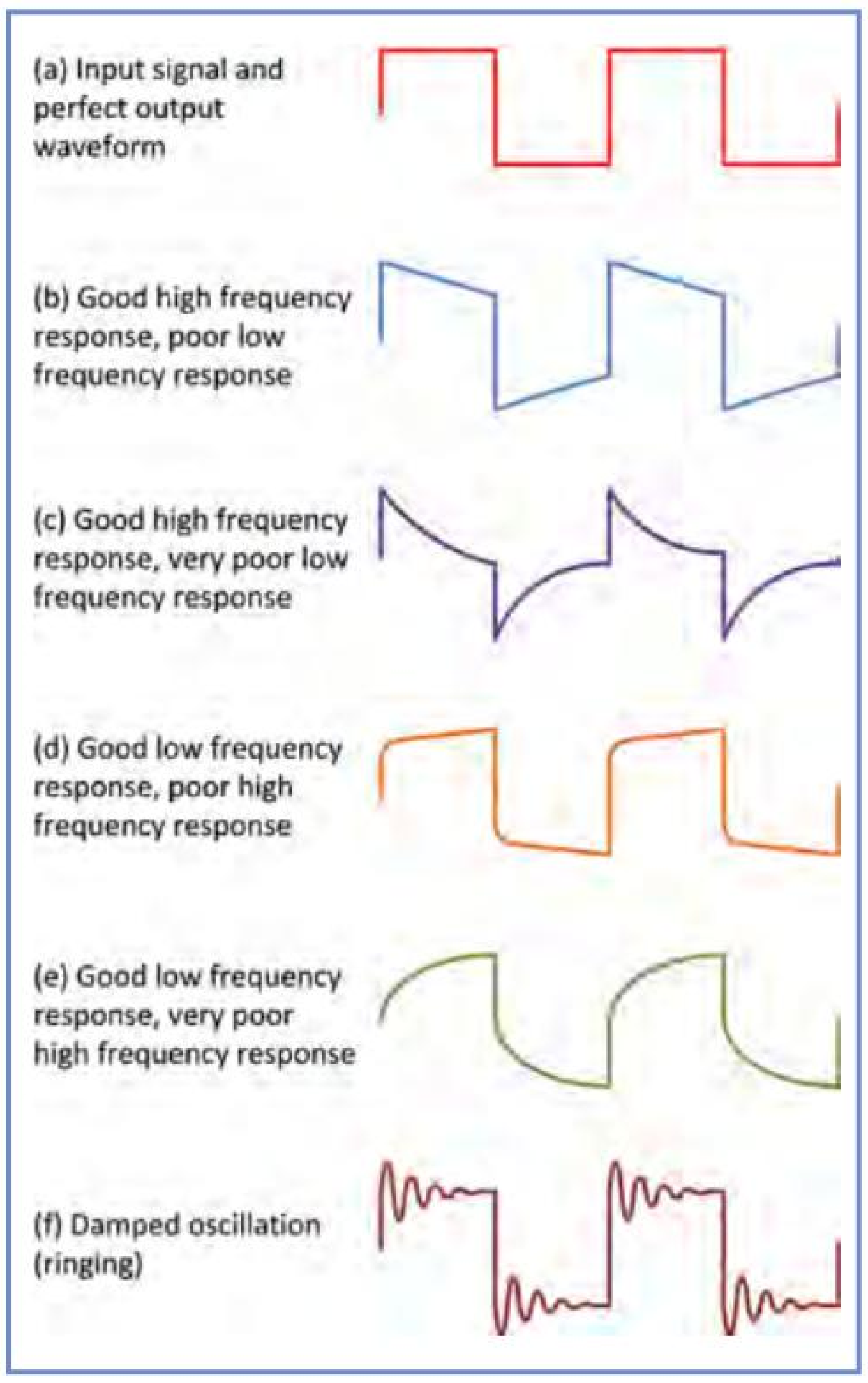

(a) Input signal and perfect output waveform
(b) Good high frequency response, poor low frequency response
(c) Good high frequency response, very poor low frequency response
(d) Good low frequency response, poor high frequency response
(e) Good low frequency response, very poor high frequency response
(f) Damped oscillation (ringing)

Fig.6.9. Square-wave testing can provide a rapid means of checking the frequency response and transient performance of an amplifier

resistance of the amplifier can then be calculated from: R_{in} = VR + R_s
Here, R_s is the source impedance of the signal generator (often 600Ω or 50Ω).

Measuring output impedance/resistance

An arrangement for measuring output resistance is shown in Fig.6.3(d). A variable resistor or decade resistance box (VR) is used as a load. This resistor is initially left disconnected (ie, open-circuit) and the amplifier is once again adjusted for normal operation (but without a load present). The output voltage is then noted, after which VR is reconnected and adjusted until the output falls to exactly half its previous value. At this point, the value of resistance is measured (or read from the decade resistance box). The output resistance (R_{out}) can then be calculated from: R_{out} = VR. Note that this method is *unsuitable* for use with power amplifiers because they have an extremely low output impedance and the excessive current demand may cause damage to the output stage.

Waveform testing

If you have a signal generator and an oscilloscope handy you can carry out a very quick check on the performance of an amplifier by simply observing the output waveform produced when a signal is applied to the input of the amplifier. The amplifier will need to have a load connected (see above) and the test signal frequency should normally be 1kHz at a level that preserves linearity and avoids any risk of over-driving the amplifier. Two different tests can be applied; one using a sinewave input and the other using a square wave. The former provides a quick check on linearity and distortion, while the latter can be used for a quick assessment of frequency response.

Sinewave testing

A quick inspection of an output waveform will usually provide you with a clue as to what type of distortion has been introduced by an amplifier, but first it is necessary to ensure that the input signal is free from distortion and this involves checking that it is a reasonably pure sinewave. Most signal generators (see page 43) are able to produce sinewave outputs with sufficiently low levels of distortion that cannot be discerned by the human eye when a waveform is viewed on an oscilloscope. As the level of distortion increases – typically to about 5% or more – the distortion starts to become visible in the form of a departure from a pure sinewave shape. This makes it possible to carry out a quick check on distortion by simply viewing the shape of the output waveform.

Fig.6.8 shows some representative test waveforms. The undistorted sinusoidal input signal is shown in Fig.6.8(a). This is also the ideal shape for the output waveform, which should be perfectly sinusoidal if the amplifier is not introducing any distortion. The effect of clipping is shown in Figs.6.8(b) to 6.8(d). In the case of Fig.6.8(b) the positive edge of the waveform has been clipped. Notice the flattening effect this has on the positive excursions of the signal. Fig.6.8(c) shows a similar effect applied to the negative edge of the waveform. These two conditions usually point to an incorrect bias adjustment where an applied signal becomes increasingly distorted (and clipped) such as whenever the amplitude of the input signal exceeds a certain value.

Symmetrical clipping (ie, clipping of both positive and negative peaks) is illustrated in Fig.6.8(d). This condition usually results from applying an input signal of excessive amplitude. Reducing the amplitude below a critical value (ie, below the point at which clipping starts to occur) will often correct the problem and reduce the distortion to an acceptable amount. Fig.6.8(e) shows cross-over distortion which can be a problem when insufficient standing current is available in a complementary push-pull output stage. Note how the output signal remains at zero until the input signal reaches a certain value and how this form of distortion affects both positive and negative-going half cycles of the input waveform.

Apart from the repetitive waveform defects that we've seen thus far, a signal might also become contaminated by random fluctuations in signal level (noise) and sudden spikes of interference caused by switching and other transients induced into wiring (both internal and external). This problem is illustrated in Fig.6.8(f).

Last, you might occasionally come across a circuit that, while intended to act as an amplifier, also acts as an oscillator! The reason for this is that, at some frequency (usually very much higher than the highest designed signal frequency) the internal phase shift becomes such that the feedback becomes positive (instead of the intended negative feedback), consequently de-stabilising the amplifier and resulting in continuous oscillation. Designers of high-gain amplifiers usually try to avoid this problem by using only local feedback (ie, feedback over a single stage). This effect is illustrated in Fig.6.8(g). Note that the amplitude of the parasitic oscillation is often significantly reduced due to the limitations of the frequency response of the oscilloscope used to observe the distorted waveform.

Square wave testing

An alternative to using a sinewave for testing an amplifier is using a square wave. This can provide you with a rapid assessment of the frequency response of an amplifier. Square waves comprise an infinite number of sinusoidal harmonic components added to the fundamental sinewave, and so any defects in the frequency response of an amplifier will show up very quickly from an examination of the shape of the waveform of the output signal when a square wave input is applied. As a result, it is possible to assess whether the frequency response is good or poor (a perfect square wave output would correspond to a perfect frequency response).

Fig.6.9 shows waveforms that correspond to several different frequency response characteristics. A perfect

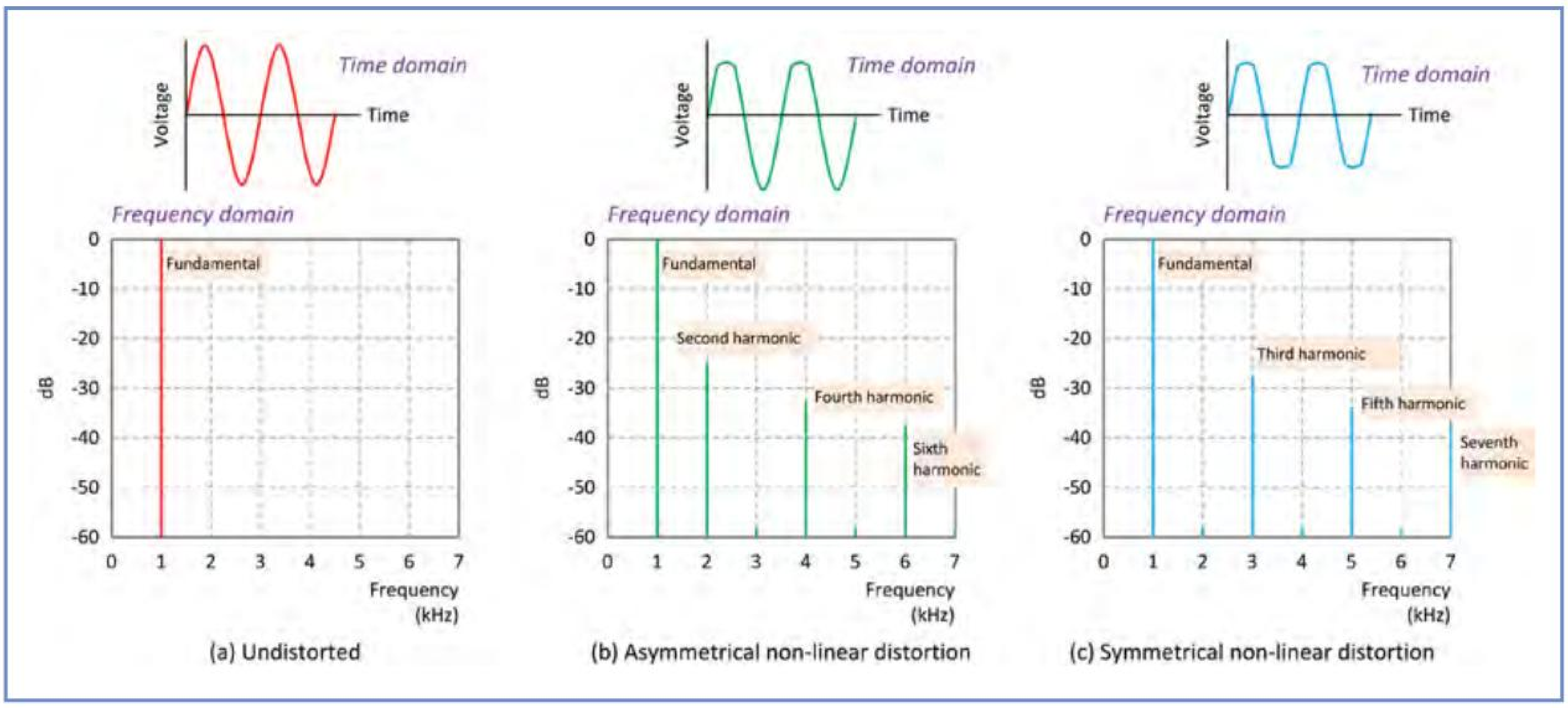


Fig.6.10. Effect of a non-linear transfer characteristic on the quality of an output waveform

square wave output, Fig.6.9(a), indicates that the amplifier under test has a flat frequency response. Figs.6.9(b) and 6.9(c) are typical of an amplifier having a good high-frequency response coupled with a poor low frequency response, while Figs.6.4(d) and 6.4(e) are indicative of poor high-frequency response and good low-frequency response.

The damped oscillation, or 'ringing', shown in Fig.6.9(f) occurs when the amplifier's step response (ie, its response to a sudden and very rapid change in signal level) produces momentary oscillation. This can occur in amplifiers where appreciable inductive reactance is present at the same time as capacitive reactance. The combined effect of these two opposite reactances results in a resonant effect, where the sudden changes imposed by the rising or falling edges of the square wave signal causes energy to oscillate back and forth between the two opposite reactive components. The transfer of energy between the two components decays due to losses (resistance) present in the circuit and so the oscillation eventually settles to a steady value at one or other extreme of amplitude.

Measuring noise and distortion

There are several different forms of distortion and all of them may, to a greater or lesser extent, be present at the same time. If an amplifier has a perfectly linear transfer characteristic (ie, voltage out plotted against voltage in) and a perfectly flat frequency response (ie, voltage out plotted against frequency) it will not produce any distortion (it might, however, be susceptible to hum and noise, as we explain later). Conversely, if the frequency response and/or transfer characteristic is imperfect then this will result in the production of distortion but this may, or may not, be a problem depending upon the severity of the non-linearity and the degree of aberration in the frequency response. Let's first consider the effect of non-linearity in the transfer characteristic.

Distortion due to transfer characteristic non-linearity

Fig.6.10(b) shows what will happen when the transfer characteristic is asymmetrical. In this case, the characteristic flattens-off beyond a particular positive-going input level. The result is that the output signal becomes prematurely truncated or 'clipped'. A similar effect would be produced if the negative-going part of the transfer characteristic had become flattened while the positive-going part remained linear. However, in this case the negative edge of the output signal would be clipped. The resulting harmonics produced will include the second, fourth and sixth, and so on, as shown in Fig.6.10(b). None of these components are present in the undistorted pure sinewave signal shown in Fig.6.10(a).

Fig.6.10(c) shows the effect of symmetrical non-linearity in the transfer characteristic. In this case, both positive and negative edges of the output waveform have become clipped. The resulting harmonics produced will include the third, fifth, seventh, and so on, as shown in Fig.6.10(c). Notice also that the output amplitude has become reduced when compared with the undistorted output waveform shown in Fig.6.10(a).

Note that an increase in the amplitude of a test signal will usually result in a significantly greater increase in the level of harmonic distortion and the amplifier becomes 'overdriven'. The problem will increase in severity with a further increase in input signal to the point that the distortion produced will very quickly reach an unacceptable level. Fig.6.11 shows the effect of overdriving a power amplifier on the amount of total harmonic distortion generated. Note the rapid increase in THD when the rated output power (8W) is exceeded.

Distortion due to aberrations in frequency response

As well as distortion resulting from non-linearity of its transfer characteristic, an amplifier will also introduce distortion due to imperfections in its frequency response. It can be a sobering experience to apply a square wave to an amplifier and find that the output waveform is not very square. The reason for this is that for a square wave to be perfectly reproduced, an amplifier would need to have a perfect frequency response. This, of course, is never actually the case.

Measuring distortion

The accurate measurement of distortion involves the use of some relatively sophisticated test equipment in the form of a spectrum analyser or wave analyser. When carrying out laboratory checks on prototypes we use a wave analyser as well as a distortion factor meter. The former instrument is capable of measuring the level of the individual harmonics present in an output signal, while the latter provides us with a figure for the circuit's THD performance in a particular bandwidth (we usually restrict our own audio measurements to an upper limit of 100kHz).

The distortion factor meter comprises a wide band (100kHz) voltmeter combined with a variable frequency notch filter that can be tuned so that it eliminates the fundamental frequency component present in the amplifier's output. In use, a reference level is set at 100 % with the filter switched out and then the level is measured again with the filter selected. In this condition, the signal measured is residual distortion present and its level (in relation to the fundamental) can be read from a panel meter in decibels (dB) or as a percentage.

It is important to be aware that, when using a distortion factor meter rather than a more complex wave analyser, the instrument will respond to all in-band signals, including noise, hum and other non-harmonically related components.

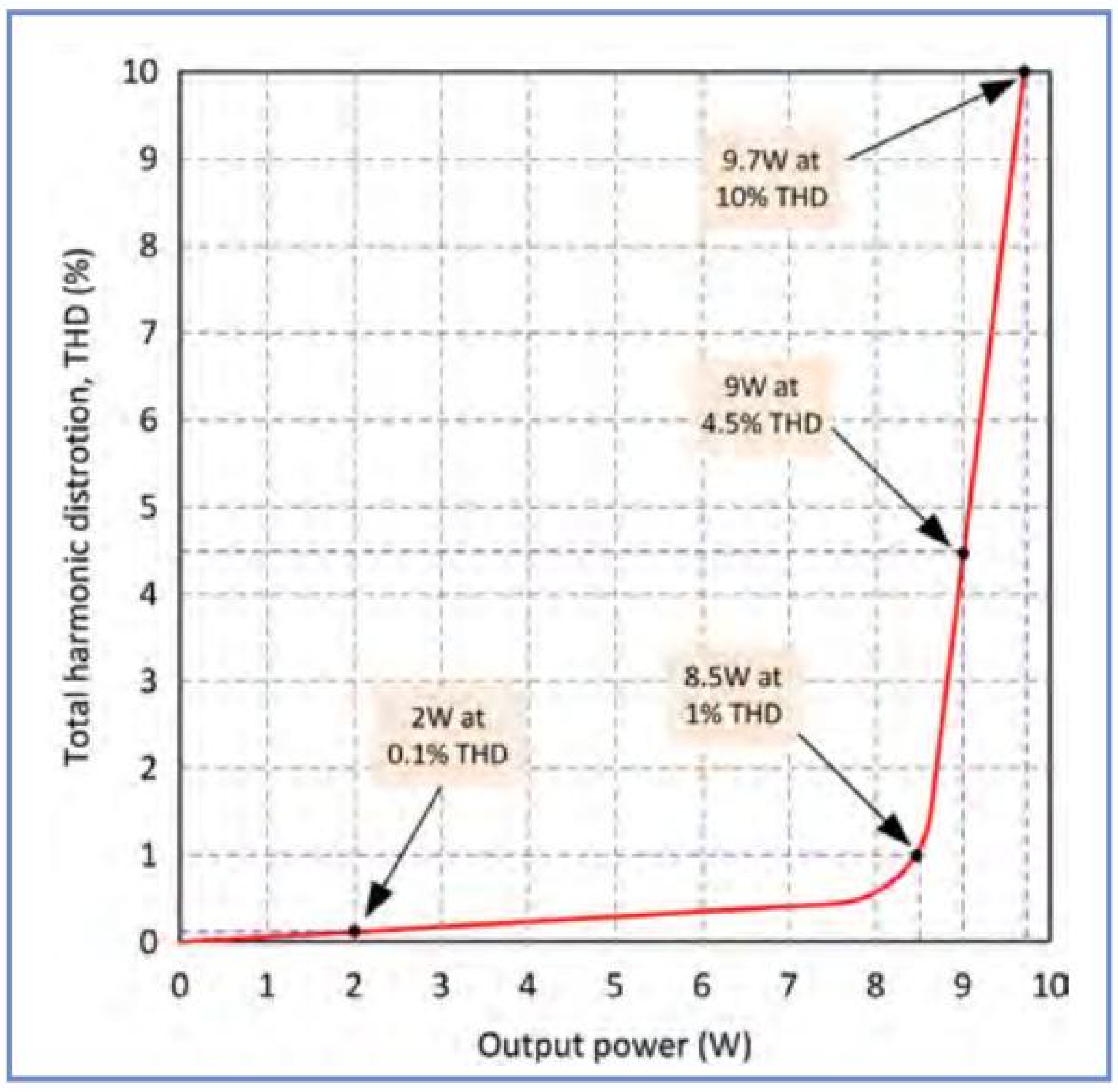


Fig.6.11. Overdriving a power amplifier results in a rapid rise in total harmonic distortion

Nevertheless, this type of instrument can be very effective when carrying out a quick assessment of the performance of an amplifier.

When working at very low levels of THD (typically less than 0.1%) the ultimate accuracy of the measurement becomes dependent on the quality of the input signal (which must be as near perfect as possible). We use a Radford low-distortion oscillator for our distortion measurements. This instrument is capable of producing a sinewave test signal with less than 0.003% and typically 0.001% THD. Note that popular low-cost function generators can often produce as much as 0.5% THD, and so this type of signal source is unsuitable for carrying out meaningful THD measurements.

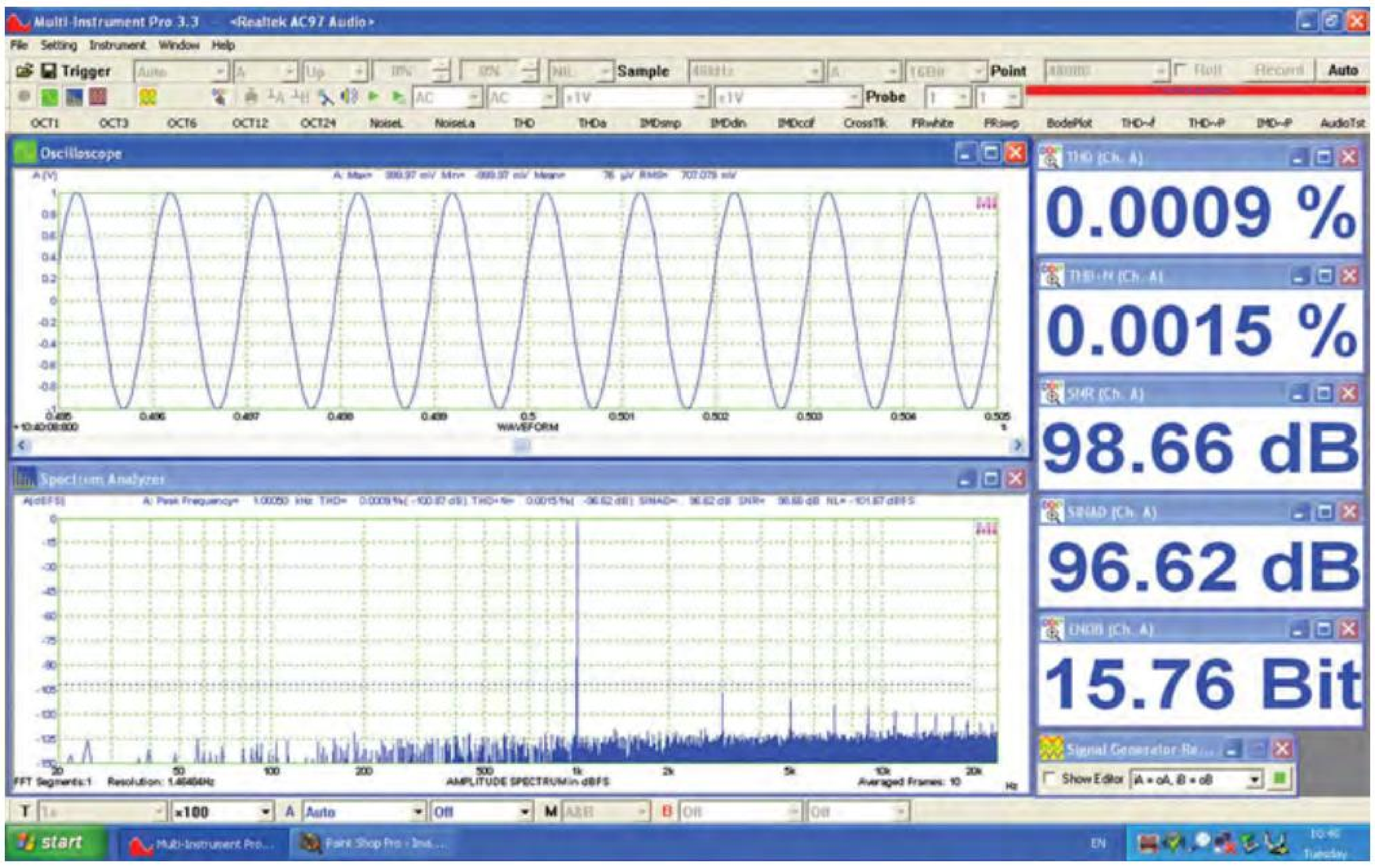
How the THD is measured is important too. Measuring THD typically requires sinewaves, but measuring THD with a signal having a level just below that which would produce clipping can be instrumental in hiding other forms of distortion, notably cross-over distortion which disproportionately affects lower signal voltages. In most situations a THD measurement done at 10dB below rated output power (eg, at 1W RMS for an amplifier rated at 10W RMS) would be a much better indicator of sound quality.

Using PC-based software

An alternative to using stand-alone test equipment for audio measurements is using one or more virtual instruments. These use PC-based software and hardware and can be used for a variety of measurements including the analysis of distortion and noise. Solutions that use external hardware connected via a USB port are usually more accurate than internal sound cards to achieve the required analogue-to-digital conversion. Nevertheless, provided that your PC has a fast sound card with a low noise floor measurements can be made that are comparable with stand-alone test equipment (and often at much lower cost).

Fig.6.12 shows a virtual instrument display being used to check a sound card output and to ensure that the THD was sufficiently low to enable accurate low-level measurements to be carried out. As you can see, a great deal of information is available from the software. The virtual oscilloscope and signal analyser windows are simultaneously displayed. Notice that the frequency scale is logarithmic and that it extends from 20Hz to 20kHz, covering the full audio frequency range. Careful examination of the frequency spectrum shows the fundamental with an amplitude of 1V, together with odd-order harmonic components with descending amplitude. This suggests that the distortion arises from symmetrical rather than asymmetric non-linearity, as discussed earlier.

Various performance figures are reported in windows on the right. The first of these is the THD figure (0.0009%).


Multi-Instrument Pro 3.3 - <Realtek AC97 Audio>
File Setting Instrument Window Help
Trigger
Sample
Point
Auto
Probe
Oscilloscope
WAVEFORM
Spectrum Analyzer
AMPLITUDE SPECTRUM in dBFS
Averaged Frames: 10
Hz
Show Editor
start

THD (Ch. A) 0.0009 %

THD+N (Ch. A) 0.0015 %

SNR (Ch. A) 98.66 dB

SINAD (Ch. A) 96.62 dB

ENOB (Ch. A) 15.76 Bit

Fig.6.12. Virtual instrument display during tests on a sound card

This is extremely low. The next figure shows THD plus noise and hum (THD+N) and this would be the same figure that a distortion factor meter would report. The THD+N figure is, of course, greater than the THD figure but it is still very acceptable at a mere 0.0015%. The third window shows the signal-to-noise ratio (SNR) and the reported value is a very acceptable 98.66dB.

The next window shows SINAD or 'signal in noise advantage' SINAD is a measure of the signal quality that takes into account the presence of distortion. SINAD is calculated from the ratio of total output signal power (ie, signal power plus noise power plus the power arising from distortion components) to the noise-plus-distortion power. As might be expected, the SINAD performance is only slightly worse than the SNR performance, but the SINAD figure is usually a more reliable measure of performance, particularly when a significant amount of noise and distortion is present in a system.

The last window shows the effective number of bits (ENOB), which provides an indication of the dynamic performance of the system in relation to that of an analogue-to-digital converter (ADC). Since the number of bits used to represent an analogue quantity is specified by the resolution of an ADC, the number displayed indicates the resolution of an ideal ADC that would operate with the same resolution as the circuit under consideration. The figure quoted here is just less than 16-bits and is indicative of a high-quality audio system.

Signal-to-noise ratio

The signal-to-noise ratio in a system is normally expressed in decibels (dB) and is defined as:

$$\text{S/N} = 10\log_{10}\left(\frac{P_{signal}}{P_{noise}}\right)\text{dB}$$

In practice, it is difficult to separate the signal present in a system from the noise. If, for example, you measure the output power produced by an amplifier you will actually be measuring the signal power together with any noise that may be present. Hence, a more practical measure is the ratio of (signal-plus-noise)-to-noise. Furthermore, provided that the noise power is very much smaller than the signal power, there will not be very much difference between the signal-to-

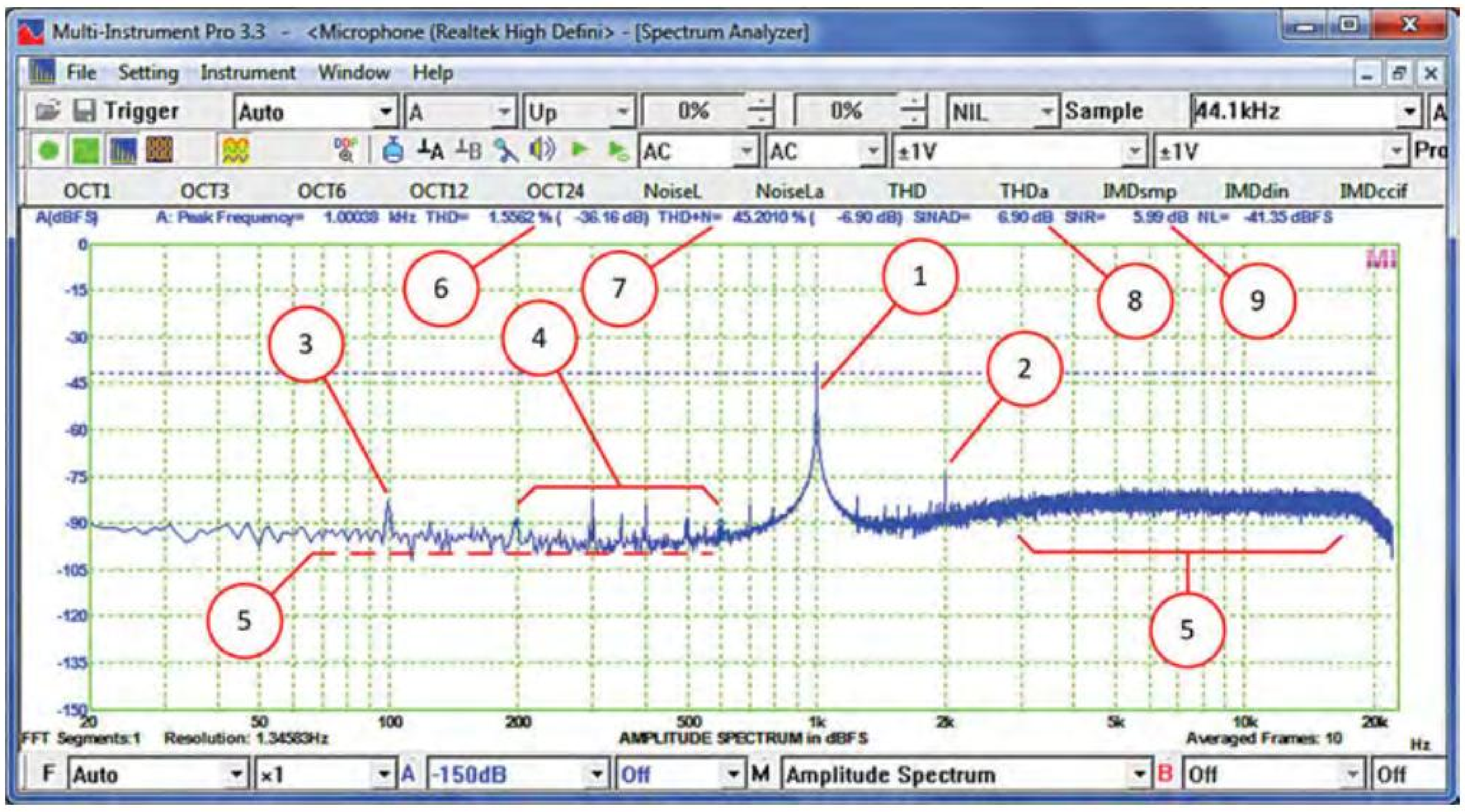

Multi-Instrument Pro 3.3 - <Microphone (Realtek High Defini> - [Spectrum Analyzer]
File Setting Instrument Window Help
Trigger
Auto
Up
0%
NIL
Sample
44.1kHz
AC
±1V
OCT1 OCT3 OCT6 OCT12 OCT24 NoiseL NoiseLa THD THDa IMDsmp IMDdin IMDccif
A(dBFS)
A: Peak Frequency= 1.00038 kHz THD= 1.5562 % (-36.16 dB) THD+N= 45.2010 % (-6.90 dB) SINAD= 6.90 dB SNR= 5.99 dB NL= -41.35 dBFS
1
2
3
4
5
6
7
8
9
0
-15
-30
-45
-60
-75
-90
-105
-120
-135
-150
20
50
100
200
500
1k
2k
5k
10k
20k
FFT Segments:1 Resolution: 1.34583Hz
AMPLITUDE SPECTRUM in dBFS
Averaged Frames: 10
Hz
Auto
x1
-150dB
Off
Amplitude Spectrum

Fig.6.13. Spectral analysis of a signal with noise and distortion present

noise-ratio and the ratio of (signal-plus-noise)-to-noise; thus:

$$(S+N)/N = 10\log_{10}\left(\frac{P_{\text{signal+noise}}}{P_{\text{noise}}}\right)\ \text{dB}$$

It might help to put this into context with some representative figures. Let's assume that, in the absence of a signal the noise power present at the output of a (rather noisy) amplifier is 100µW and when the signal is applied the output power increases to 400mW. The (signal-plus-noise)-to-noise ratio can be calculated from:

$$(S+N)/N = 10\log_{10}\left(\frac{400\text{mW}}{100\mu\text{W}}\right)$$

$$= 10\log_{10}(4000) = 10 \times 3.6 = 36\text{dB}$$

The (signal-plus-noise)-to-noise ratio may also be determined from the voltages produced by an amplifier, in which case:

$$(S+N)/N = 20\log_{10}\left(\frac{V_{\text{signal+noise}}}{V_{\text{noise}}}\right)\ \text{dB}$$

Spectral analysis – measuring signals in the presence of noise and distortion

Spectral analysis of a signal can be extremely useful in a wide range of practical situations and it can become invaluable when dealing with noise and distortion. As an example, the frequency spectrum of a 1kHz sinewave signal is shown in Fig.6.13 where there is appreciable levels of noise, hum and distortion present. The display was obtained using the Virtins Multi-Instrument Fast Fourier Transfer (FFT) software (see *Gearing up*). If you take a careful look at Fig.6.13 you should be able to recognise the following:

1. The fundamental of the wanted signal at 1kHz (with a level of about –40dB)
2. The second harmonic of the wanted signal at 2kHz with a level of about –75dB (35dB lower than the fundamental)
3. A component at 100Hz (twice the mains supply frequency) with an amplitude of about –82dB. This was caused by a small amount of ripple present on the amplifier's DC supply
4. Harmonics of the supply ripple at 200Hz, 300Hz, 400Hz...
5. A noise floor of about –97dB with a slight increase in noise between about 2kHz and 20kHz
6. A reported THD of 1.5562% (–36.16dB relative to the 1kHz fundamental)
7. A reported THD plus noise (THD+N) of approximately 45% (–6.9dB relative to the 1kHz fundamental) – contrast this with the THD figure without noise!
8. A reported SINAD figure of 6.9dB
9. A reported signal-to-noise ratio of 5.99dB (unacceptably low for most applications).

This example gives you an appreciation of just how useful spectral analysis can be when signals are contaminated with both noise and distortion.

Gearing up: wideband AC range extender

For general audio measurements you will need, as a minimum, a good quality sinewave signal source together with an oscilloscope and an AC millivoltmeter (both described earlier *Teach-In 2018*). In addition, a dedicated AF power meter and a THD analyser would also be useful. With the exception of a power meter (which will normally incorporate a suitably rated load) these instruments can also be realised using computer-based software and hardware.

Audio frequency signal generators

An audio frequency (AF) signal generator, 'audio oscillator', or 'RC oscillator', can be a very useful investment if you are planning to carry out measurements on a regular basis. A typical AF signal generator will be capable of providing a good quality sinewave output over a frequency range of at least 10Hz to 20kHz, and ideally higher. The equipment should be fitted with a calibrated adjustable attenuator so that output levels ranging from at least 1mV to 1V can be applied to the equipment on test.

For a new instrument you can expect to pay around £100 but second-hand AF signal generators are frequently available at bargain prices from as little as £20 to £50. Instruments by Farnell, Levell, Gould, Advance and Marconi are regularly available. You might be tempted to invest in a modern direct digital synthesis (DDS) waveform generator. Such instruments sell new for around £80 but, although they might appear to have an excellent specification in terms of frequency range and waveform capability, they invariably produce an output signal that just isn't good enough for THD measurement. A purpose-designed AF signal generator should be capable of producing an output with less than 0.01% THD. This is vastly better than the '≤0.8%' offered by low-cost entry-level DDS instruments. This same reservation applies to cheaper 'function generators' based on waveform generator chips where the THD can be greater than 1%. By comparison, the author's Radford LDO3 low-distortion oscillator achieves a typical THD of 1,000-times better at a mere 0.001%. However, if you are not concerned with measuring distortion a general-purpose signal or waveform generator will be adequate for you needs.

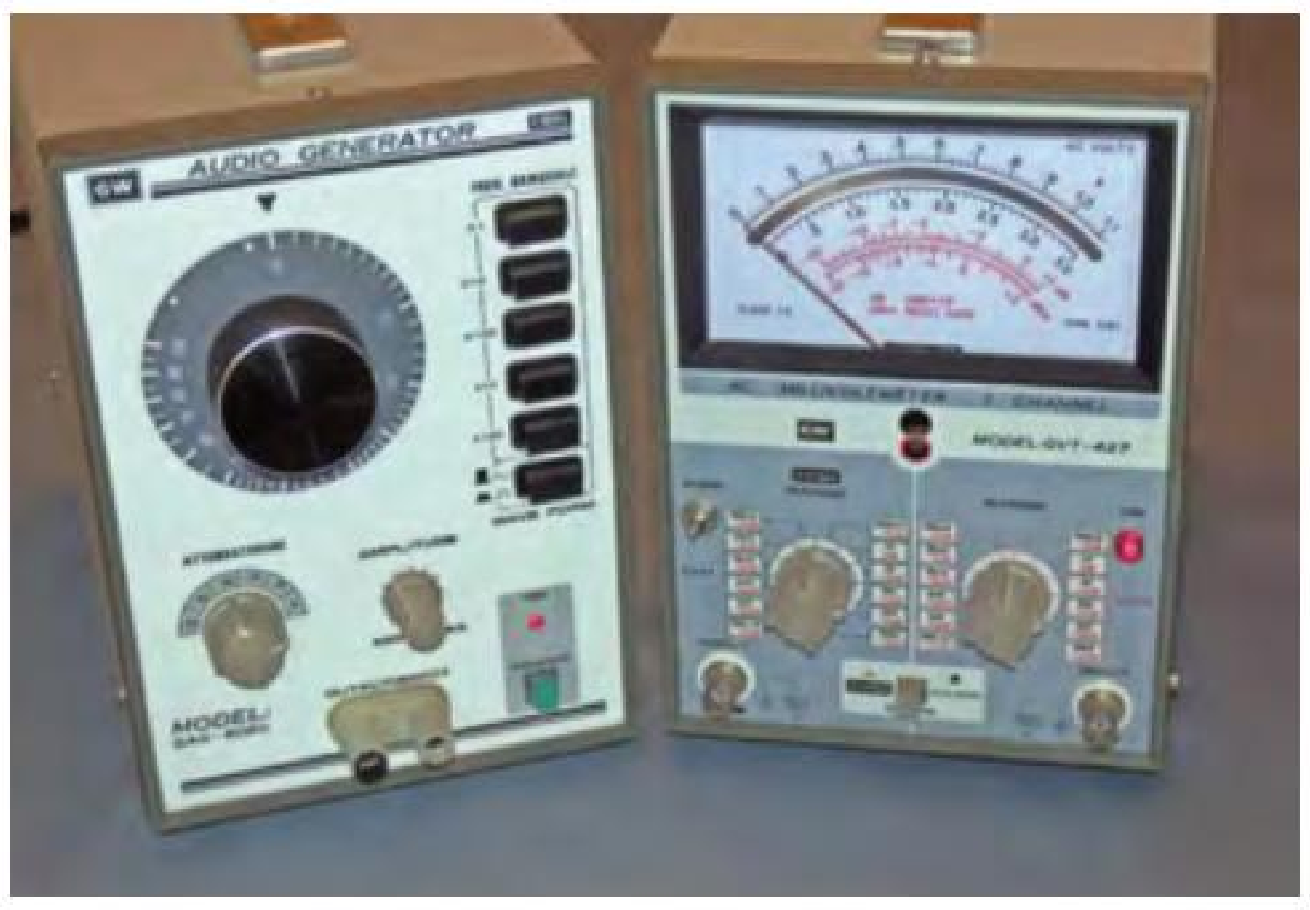


Fig.6.14. An audio signal generator (left) and millivoltmeter (right)

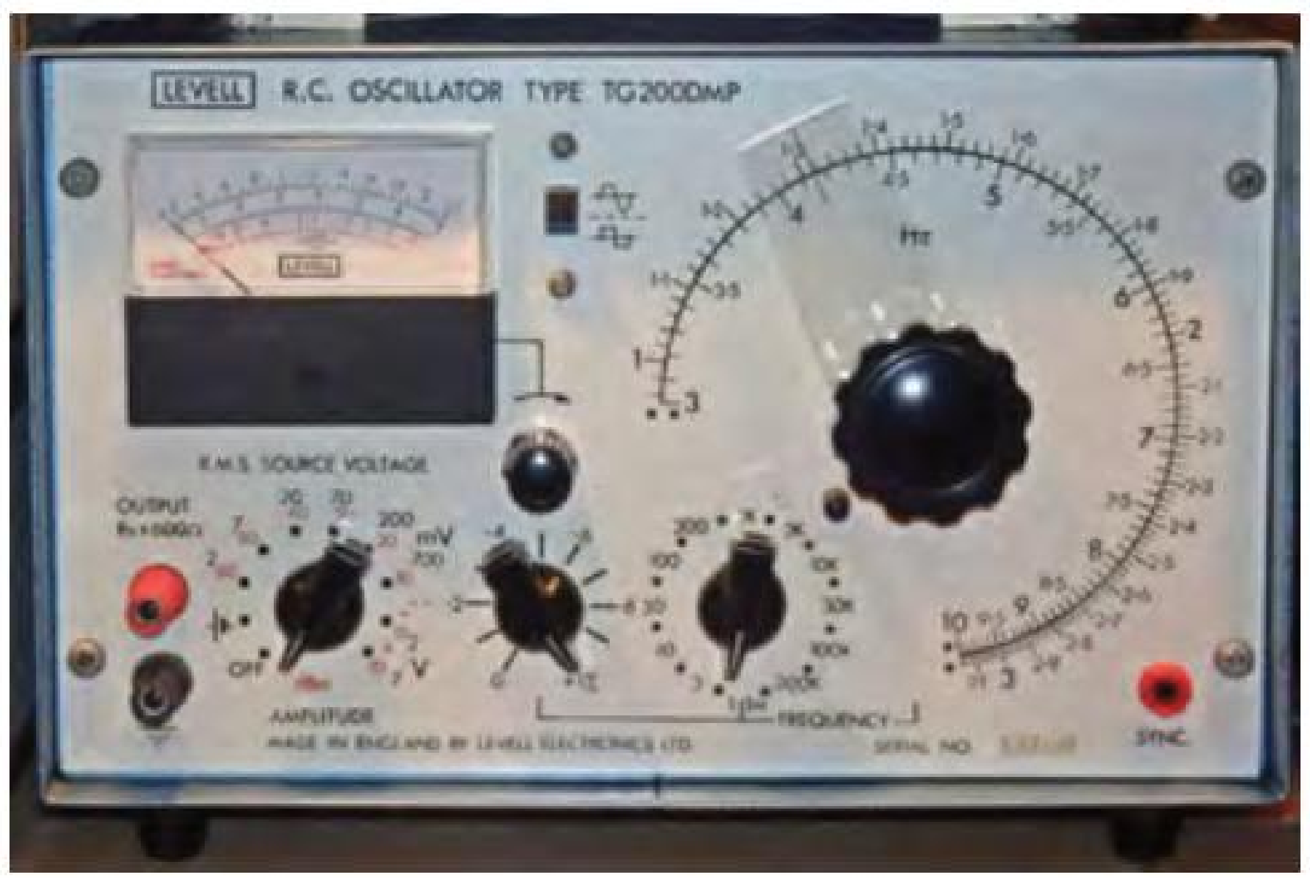


Fig.6.15. An RC oscillator with sine and square wave outputs in the range 1Hz to 3MHz, and a THD of less than 0.05% at 1kHz. Similar instruments are regularly available on-line

It's perhaps also worth mentioning that a simple AF signal generator can make an excellent project for home construction and the complete circuit of a simple Wien bridge audio oscillator is shown in Fig.6.16. This particular design uses two low-cost operational amplifiers and produces variable outputs adjustable from less than 1mV to 1V RMS with four switched decade frequency ranges extending from 2Hz to 20kHz.

AF power meters

As mentioned earlier, a simple method for measuring audio power can be based

Get it right when using an AC voltmeter

- Always ensure that test signals are free from hum, noise and distortion. This is particularly important when carrying out THD measurements.
- Avoid over-driving amplifiers and filters and always keep input signals within the normal range for the equipment on-test
- When carrying out input/output impedance and gain measurements always ensure that the test frequency is set to the centre of the mid-band of an amplifier (or to the middle of the pass-band for a filter)
- Always check that the correct load impedance is used when carrying out audio power measurements and that the load is rated for continuous operation (if this isn't the case you may need to de-rate the load or conduct tests for a short time only)
- If you are using a virtual instrument based on a PC-based sound card it is well worth checking that you are using the full capability of your sound card (software settings often default to lower-than-optimum bit rates)
- Don't rely on measurements where THD and SINAD indications may be towards the end of the instrument's measuring range (accuracy will invariably be impaired as an instrument's limits are approached).

on nothing more than a suitably rated load resistor and an accurate AC voltmeter. However, this method requires calculation and can be a little tedious when a large number of measurements are required. Instead, a dedicated audio wattmeter can provide a much more convenient solution. Unfortunately, such instruments seem to be in limited supply, but the vintage Heathkit AW-1U is occasionally available from on-line auction sites and can often be obtained for around £20 to £50. This instrument (see Fig.6.17) can read power to ±1dB over the frequency range 10Hz to 250kHz and power levels of up to 50W (25W continuous) with internal loads of 3Ω, 8Ω, 15Ω and 600Ω. Marconi TF893 and TF2500 audio output meters are also available from time to time and they can be an excellent investment.

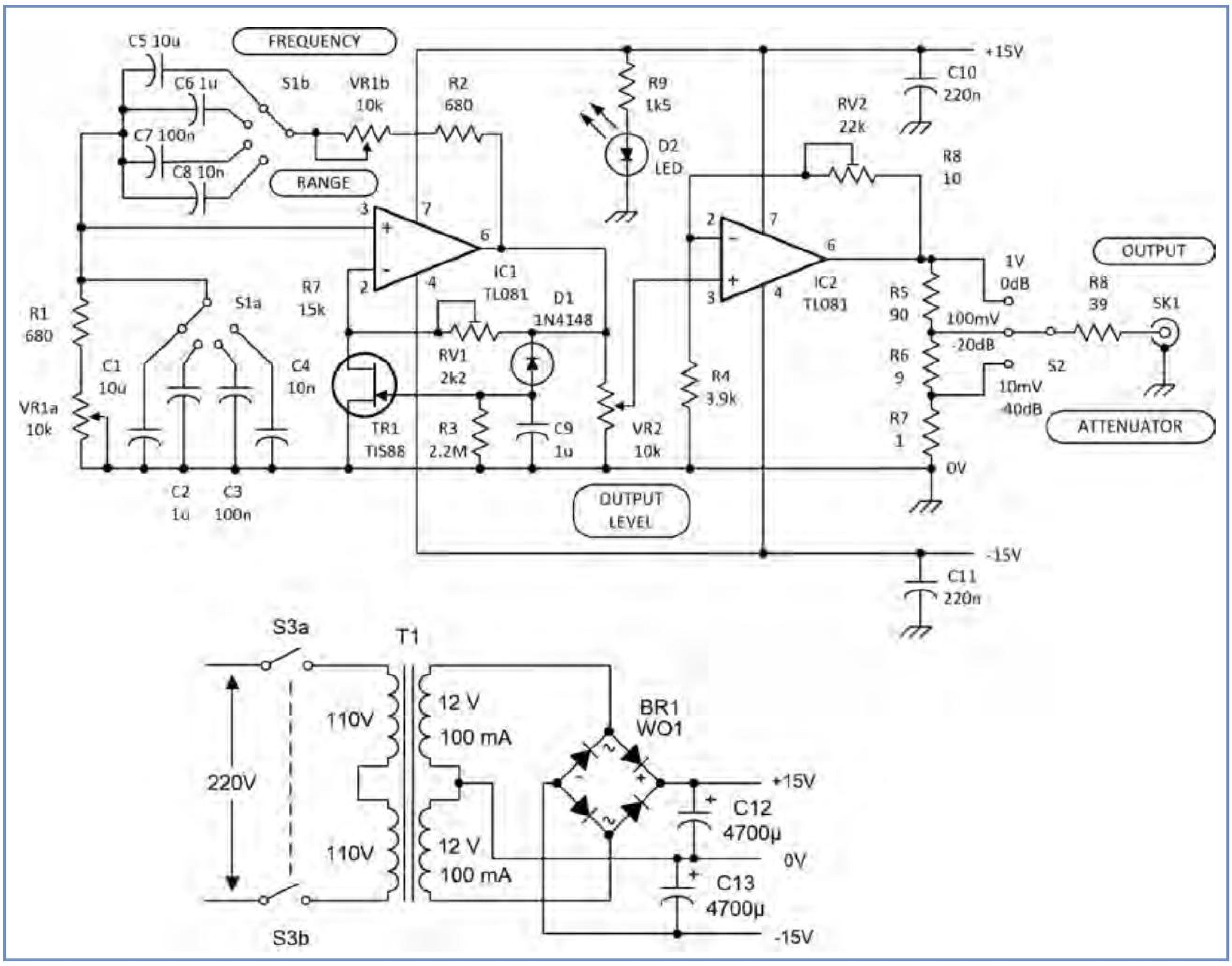

C5 10u
FREQUENCY
+15V
C6 1u
S1b
VR1b 10k
R2 680
R9 1k5
C10 220n
C7 100n
RV2 22k
C8 10n
D2 LED
RANGE
R8 10
3
7
6
2
4
IC1 TL081
R1 680
S1a
R7 15k
D1 1N4148
2
7
6
3
4
IC2 TL081
1V 0dB
OUTPUT
R8 39
SK1
R5 90
100mV -20dB
C1 10u
C4 10n
RV1 2k2
R6 9
S2
R4 3.9k
10mV -40dB
VR1a 10k
TR1 TIS88
R3 2.2M
C9 1u
VR2 10k
R7 1
ATTENUATOR
0V
C2 1u
C3 100n
OUTPUT LEVEL
-15V
C11 220n
S3a
T1
220V
110V
12 V 100 mA
BR1 WO1
+15V
C12 4700µ
0V
110V
12 V 100 mA
C13 4700µ
-15V
S3b

Fig.6.16. Complete circuit of a simple AF signal generator suitable for home construction

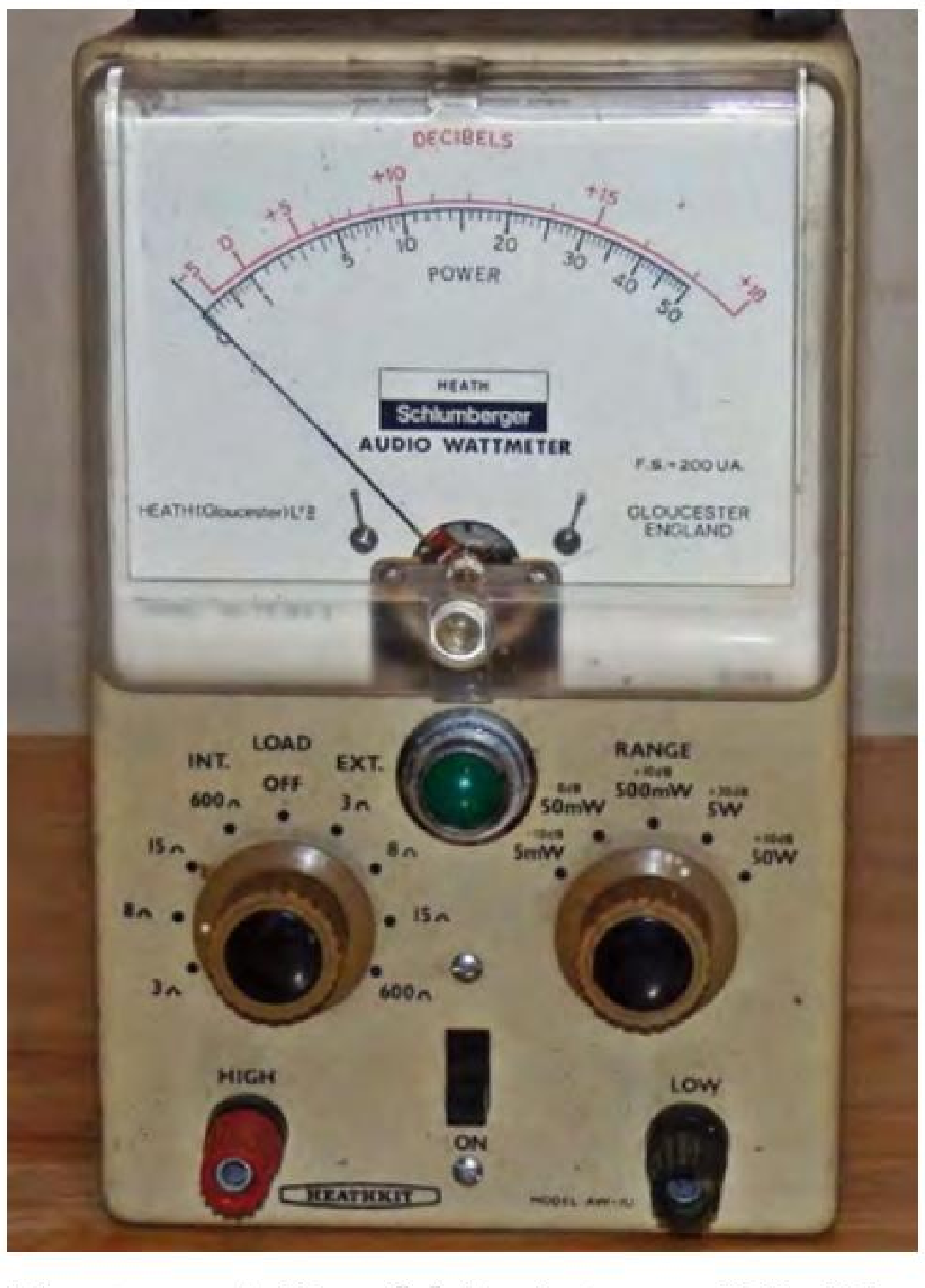

DECIBELS
POWER
HEATH
Schlumberger
AUDIO WATTMETER
GLOUCESTER ENGLAND
LOAD
INT.
EXT.
OFF
RANGE
HIGH
LOW
ON
HEATHKIT

Fig.6.17. A Heathkit vintage AW-1U audio output power meter

Distortion analysers

Distortion analysers can be expensive and, as mentioned earlier, have been replaced by powerful PC-based FFT software and fast ADC hardware. However, instruments from HP/Agilent, Leader, Marconi and Keithley do become available from time to time at prices ranging from around £50 to over £500. The author's own Marconi TF2331 distortion analyser is shown in Fig.6.18. This instrument functions as both an analyser and wideband AC voltmeter and it can measure THD down to 0.01%. The TF2331's range switch and meter calibration is shown in Fig.6.19.

Phase meters

Phase angle can be measured with a limited accuracy using a dual-channel oscilloscope, but for accuracy and convenience a dedicated phase meter is invariably a better choice. It will provide a means of measuring the phase angle

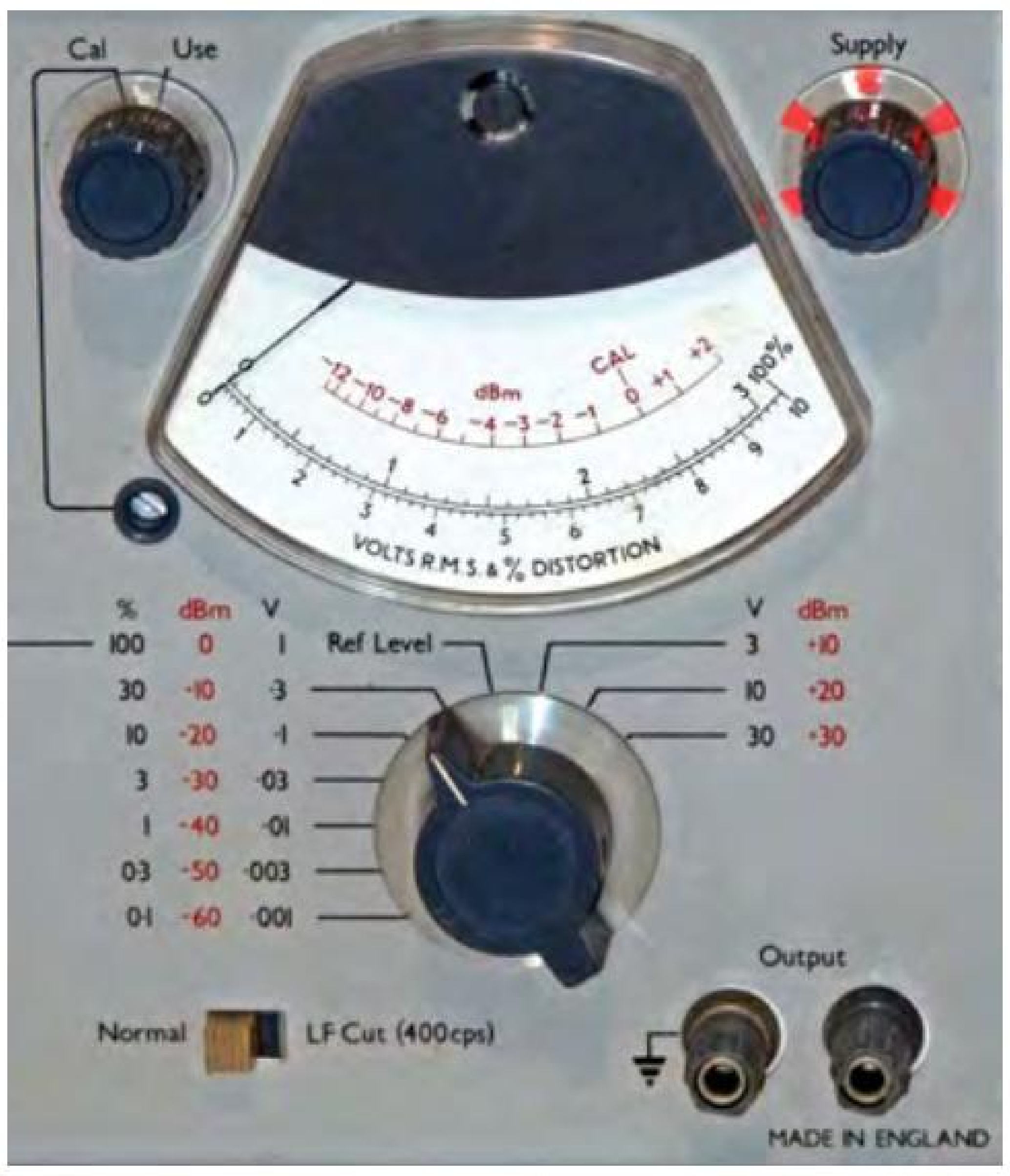

Cal
Use
Supply
dBm
CAL
VOLTS R.M.S. & % DISTORTION
Ref Level
Output
Normal
LF Cut (400cps)
MADE IN ENGLAND

Fig.6.19. Range switch and meter calibration of the TF2331 distortion analyser (Fig.6.18)

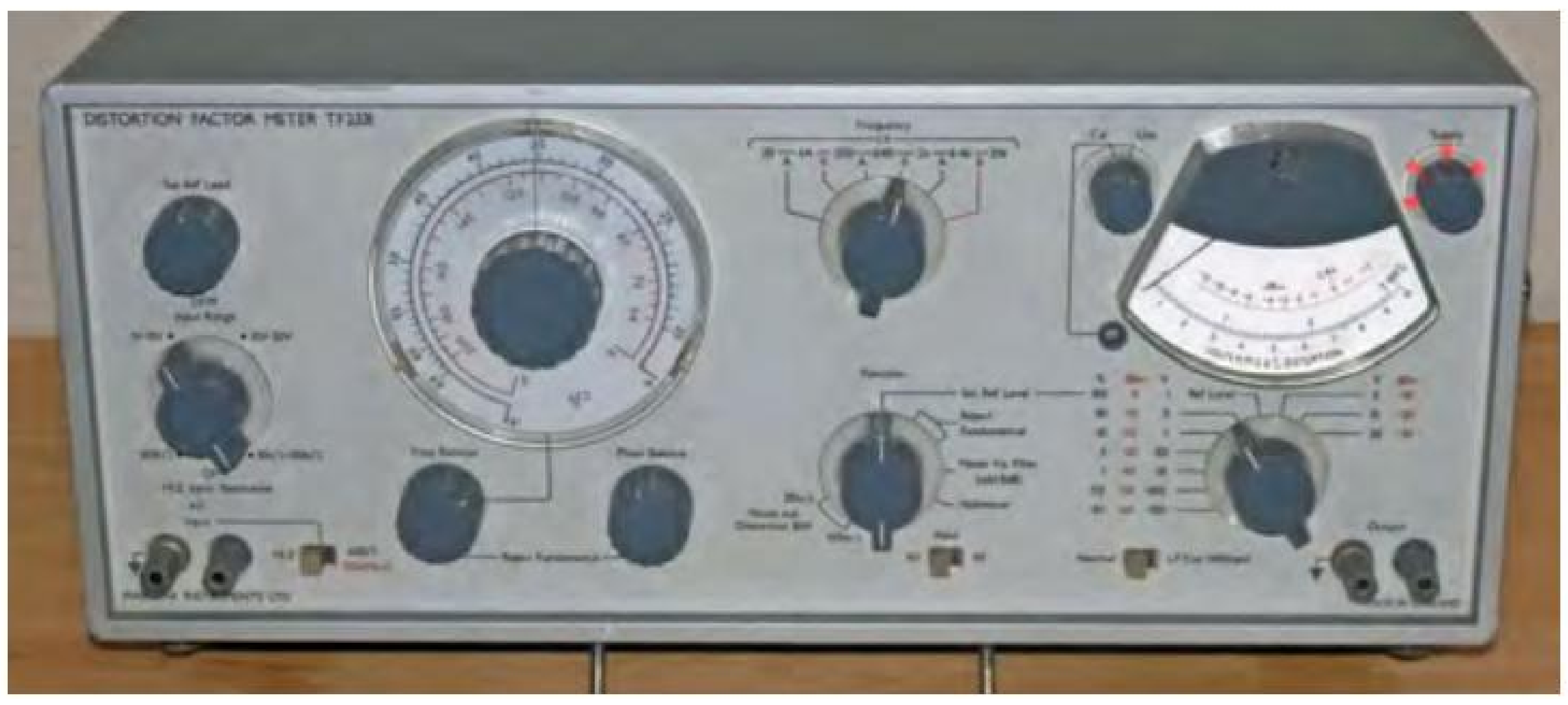

DISTORTION FACTOR METER TF2331

Fig.6.18. The Marconi TF2331 distortion analyser

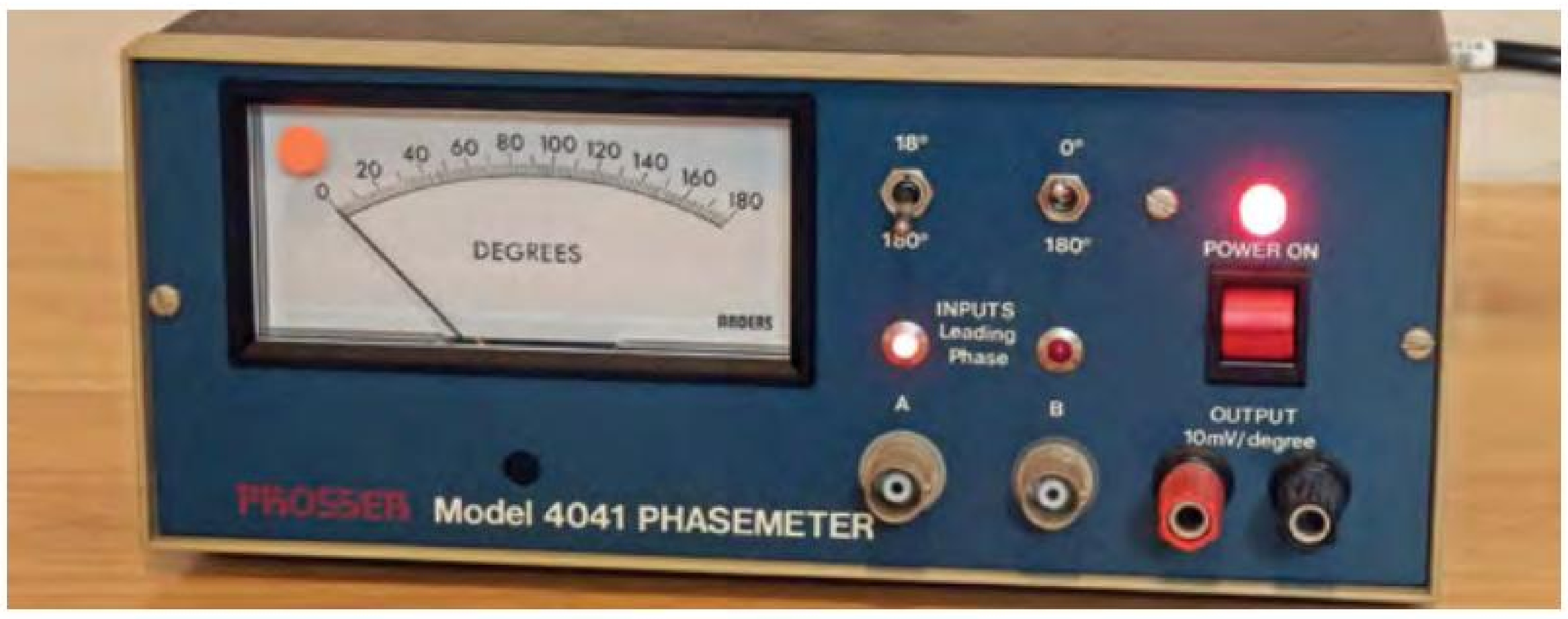

DEGREES
18°
0°
180°
POWER ON
INPUTS Leading Phase
A
B
OUTPUT 10mV/degree
PROSSER Model 4041 PHASEMETER

Fig.6.20. Phase meter with analogue display. The scale is calibrated from 0° to 180°, and the sign of the phase relationship (either leading or lagging) is indicated by a panel LED

between two waveforms (eg, the input and output of an amplifier). It is usually displayed on a scale (either analogue or digital) with a range extending from 0° to ±180° or 0° to 360°. Note that conventional phase meters can often be erratic when small levels of noise and distortion are present, and the result is often incorrect and unstable indications. Modern phase meters overcome this by using Discrete Fourier Analysis (DFT) to reject any noise and distortion without the need for tracking filters.

Computer-based virtual instruments

In *Part 2*, we introduced software and hardware packages designed primarily for use as virtual oscilloscopes. The software supplied with these instruments usually has FFT capability and so it can also be a valuable tool for distortion analysis. So, if you have a reasonably fast PC with a good quality sound card you will be able to use these for a variety of audio measurements, not just observing waveforms. If you need a wider range of features and greater accuracy then it is worth considering PC instruments from Pico Technology, Virtins, Hantek and many others.

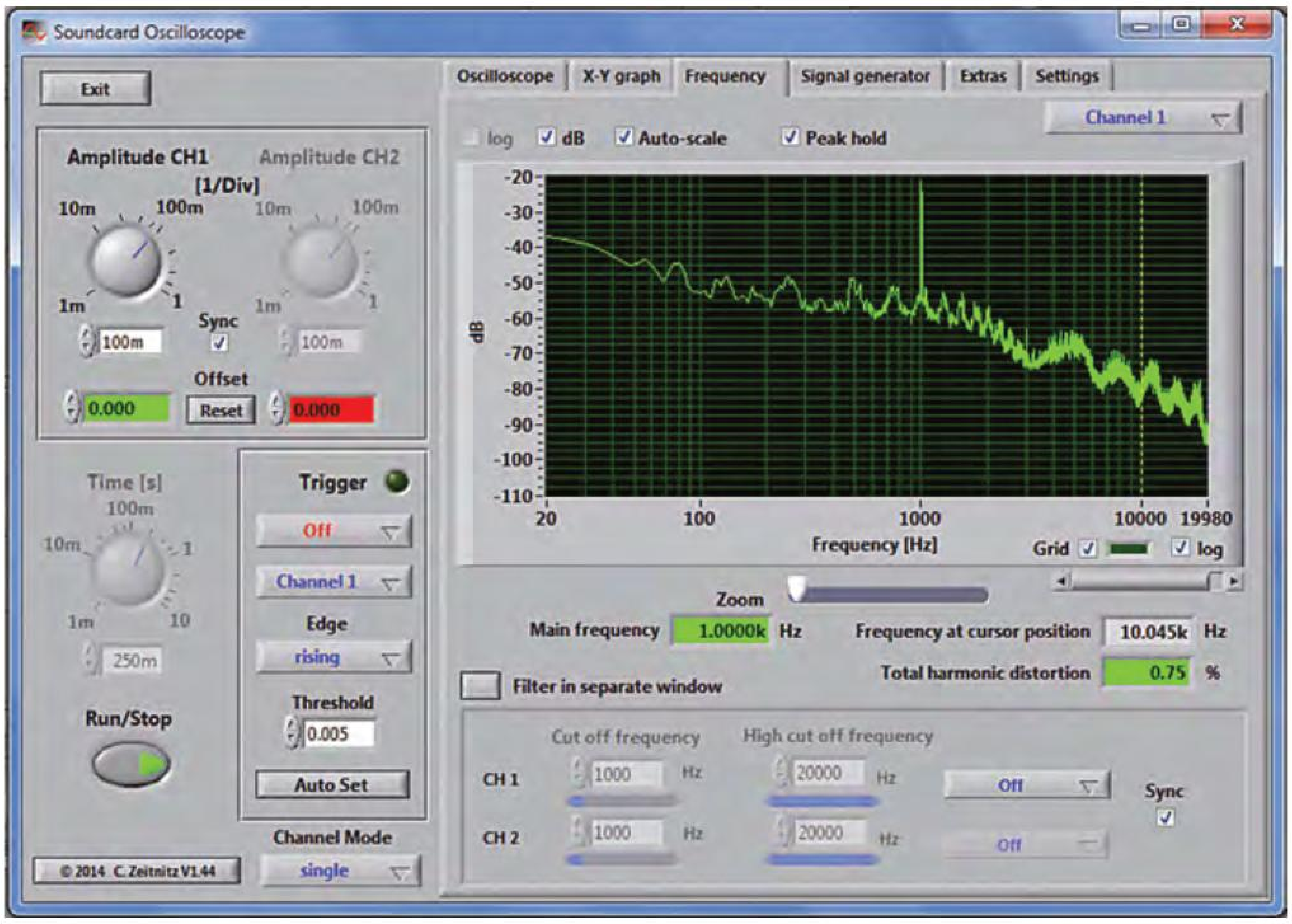
Test Gear Project: A handy test signal source

Our handy test signal source will provide you with a low-distortion 1kHz sinewave signal that will allow you to test a wide variety of audio circuits. It can be used for sensitivity, voltage gain, and input/output impedance measurements, as well as carrying out quick waveform checks for distortion.

The complete circuit of our *Test gear project* is shown in Fig.6.23. The circuit comprises an oscillator based on a twin-T phase-shift network. The frequency of oscillation is determined by the component values used in the feedback network. R2, R3 and C4 form one branch; and C2, C3 and R4 (in series with the fine frequency adjustment, RV1) form the other branch of the twin-T network. The gain of the amplifier stage (TR1) is made adjustable by means of RV2. The output amplitude is made adjustable by means of VR1 and the output signal (approximately 150mV RMS) is made available at the two front-panel-mounted 2mm sockets, SK1 and SK2. With careful adjustment the output signal THD can be as little as 1%.

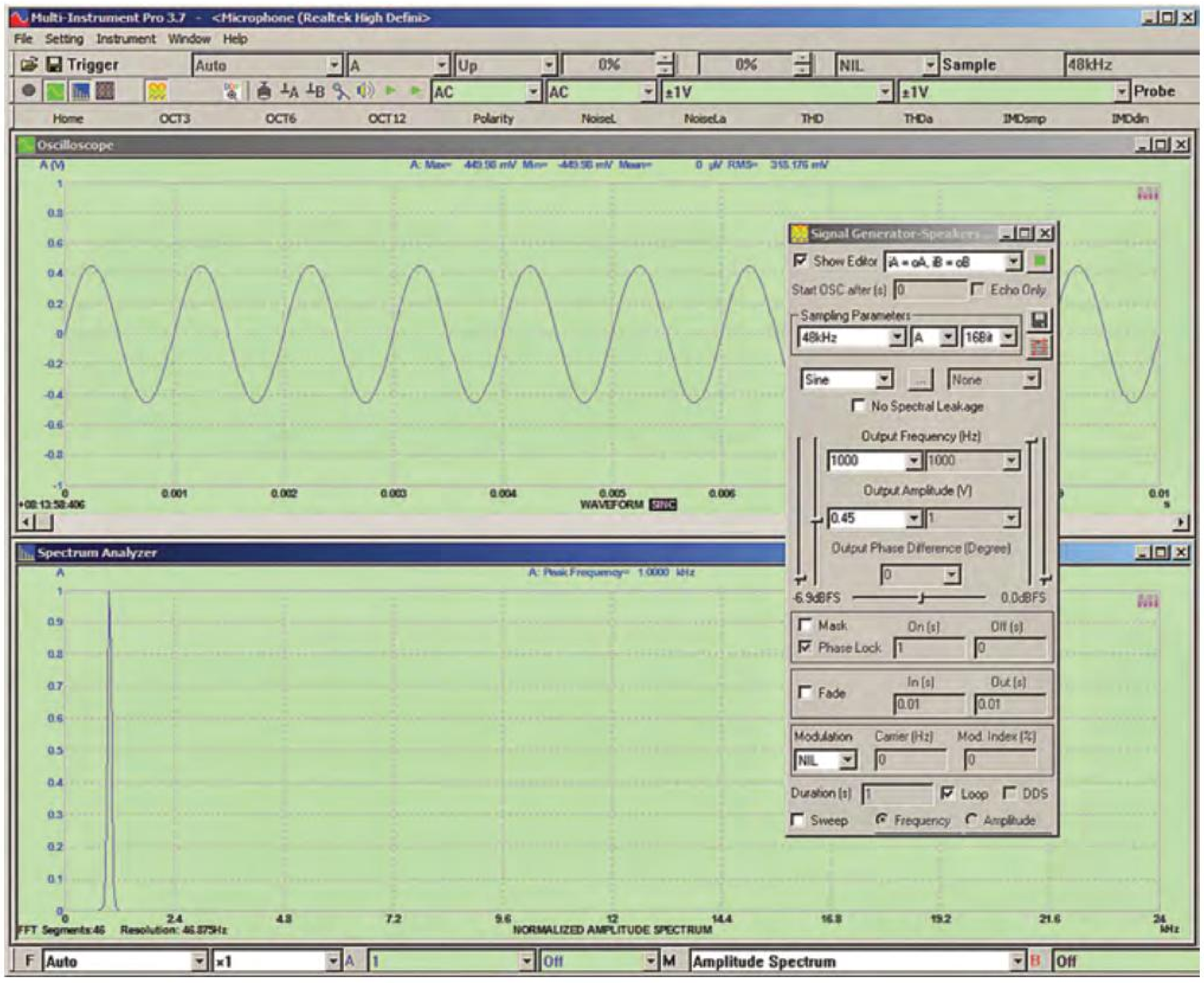
You will need

Perforated copper stripboard (9 strips, each with 25 holes)
ABS case with integral battery compartment
9V PP3 battery clip
9V PP3 battery
Miniature DPDT toggle switch (S1)
2 2-way miniature terminal blocks (ST1 and ST2)
1 red 2mm panel-mounting socket (SK1)


Soundcard Oscilloscope
Exit
Oscilloscope
X-Y graph
Frequency
Signal generator
Extras
Settings
Amplitude CH1
Amplitude CH2
[1/Div]
Sync
Offset
Reset
Time [s]
Trigger
Run/Stop
Auto Set
Channel Mode
© 2014 C.Zeitnitz V1.44
dB
Frequency [Hz]
Grid
Zoom
Filter in separate window
Cut off frequency
High cut off frequency
CH 1
CH 2

☐ log ☑ dB ☑ Auto-scale ☑ Peak hold
Channel 1
Amplitude CH1: 100m
Offset: 0.000
Time [s]: 250m
Trigger: Off
Channel 1
Edge: rising
Threshold: 0.005
Channel Mode: single
Main frequency 1.0000k Hz
Frequency at cursor position 10.045k Hz
Total harmonic distortion 0.75 %
CH 1: 1000 Hz, 20000 Hz, Off
CH 2: 1000 Hz, 20000 Hz, Off
Sync ☑

Fig.6.21. A typical frequency domain display produced by Christian Zeitnitz's software. it works with an internal PC sound card and it also provides you with a handy audio signal generator


Multi-Instrument Pro 3.7 - <Microphone (Realtek High Defini>
File Setting Instrument Window Help
Trigger
Sample
Probe
Home OCT3 OCT6 OCT12 Polarity NoiseL NoiseLa THD THDa IMDsmp IMDdn
Oscilloscope
WAVEFORM
Signal Generator-Speakers
Spectrum Analyzer
NORMALIZED AMPLITUDE SPECTRUM
Amplitude Spectrum

Sample: 48kHz
☑ Show Editor
Sampling Parameters: 48kHz, 16Bit
Sine
☐ No Spectral Leakage
Output Frequency (Hz): 1000, 1000
Output Amplitude (V): 0.45, 1
Output Phase Difference (Degree): 0
☐ Mask
☑ Phase Lock
☐ Fade
Modulation: NIL
Duration (s): 1 ☑ Loop ☐ DDS
☐ Sweep ● Frequency ○ Amplitude
A: Peak Frequency= 1.0000 kHz

Fig.6.22. The Virtins Multi-Instrument signal generator producing a 1kHz sinewave with an amplitude of 450mV. Digital loopback has been enabled so that the waveform and frequency spectrum can be concurrently displayed in the oscilloscope and spectrum analyser windows

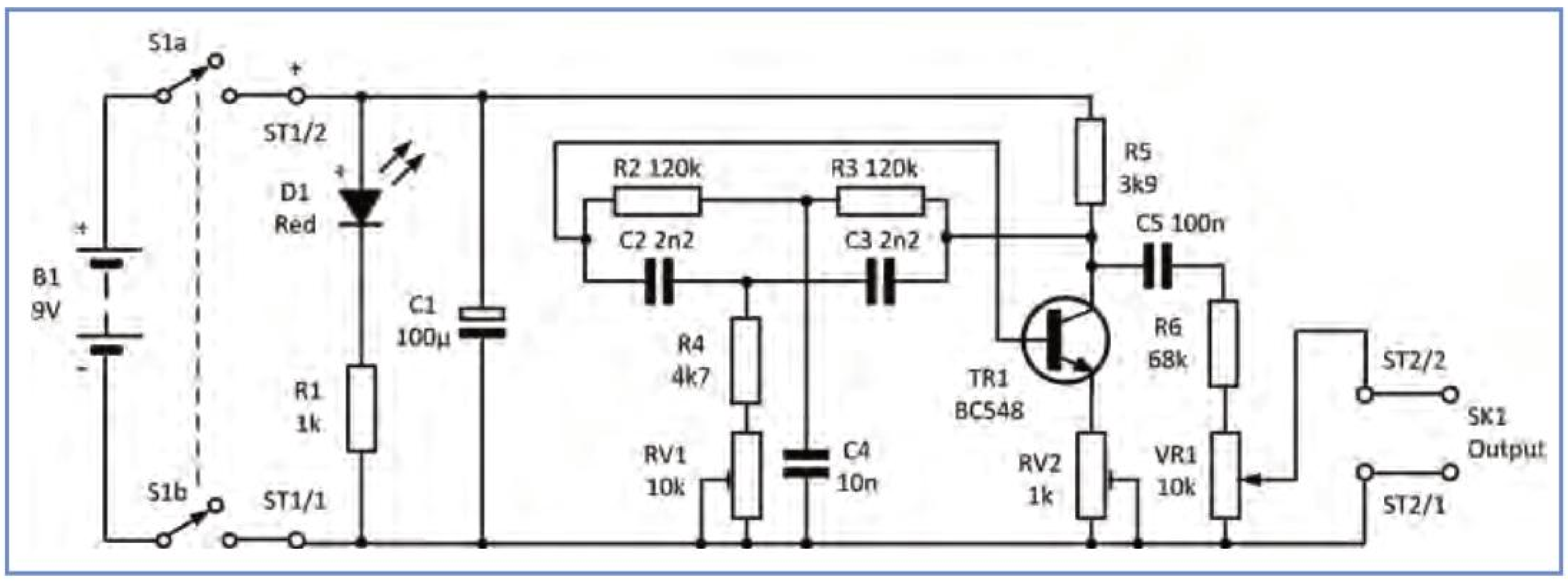

S1a
ST1/2
D1
Red
B1
9V
C1
100µ
R1
1k
S1b
ST1/1
R2 120k
R3 120k
C2 2n2
C3 2n2
R4
4k7
RV1
10k
C4
10n
R5
3k9
C5 100n
TR1
BC548
R6
68k
RV2
1k
VR1
10k
ST2/2
ST2/1
SK1
Output

Fig.6.23. Complete circuit of the handy test signal source

1 black 2mm panel-mounting socket (SK2)
1 BC548 transistor (TR1)
1 5mm red LED (D1)
1 1kΩ resistor (R1)
2 120kΩ resistors (R2 and R3)
1 4.7kΩ resistor (R4)
1 3.9kΩ resistor (R5)
1 68kΩ resistor (R6)
1 10kΩ miniature multi-turn pre-set resistor (RV1)
1 1kΩ miniature multi-turn pre-set resistor (RV2)
1 10kΩ linear variable potentiometer (VR1)
1 100µ 16V radial electrolytic (C1)
2 2.2nF disc ceramic capacitors (C2 and C3)
1 10nF disc ceramic capacitors (C4)
1 100nF disc ceramic capacitors (C5)

Assembly is straightforward and should follow the component layout shown in Fig.6.24. Note that the '+' symbol shown on D1 indicates the more-positive (anode) terminal of the LED. The pin connections for the LED and transistor are shown in Fig.6.25. The reverse side of the board (*not* an X-ray view) is also shown in Fig.6.24. Note that there's a total of 23 track breaks to be made. These can be made either with a purpose-designed spot-face cutter or using a small drill bit of appropriate size. There are also eight links that can be made with tinned copper wire of a suitable diameter or gauge (eg, 0.6mm/24SWG). When soldering has been completed it is very important to carry out a careful visual check of the board, as well as an examination of the track side of the board, looking for solder splashes and unwanted links between tracks. The internal and rear panel wiring of the test signal source is shown in Fig.6.26.

Setting up

Setting up is reasonably straightforward, but ideally you will need an oscilloscope and a digital frequency meter to calibrate the source and ensure that the output signal is undistorted. It is still possible to set the circuit up without these test instruments. Connect the oscilloscope to the output (if you don't have an oscilloscope you can just connect the output to an amplifier and use a speaker to monitor the signal). Next, set RV2 to minimum (zero resistance between TR1 emitter and 0V) and switch on. The output waveform (along with some noticeable distortion) should then be displayed on the 'scope (or heard from the loudspeaker).

Slowly increase RV2 until the oscillation stops, then back off the adjustment until it just starts again. At this point the output waveform will be a reasonably pure sinewave with an amplitude of about 150mV. Next, switch off momentarily and then switch on again. Check that oscillation restarts. If not, repeat the process but back off the setting of RV2 a little further until oscillation starts reliably. If you have a digital frequency meter available this can be connected to the output and RV1 can be adjusted for an output of exactly 1kHz.

If you only have an oscilloscope available you can set RV1 to produce an output waveform with a period of exactly 1ms. This will be less accurate than using a digital frequency meter, but you should still be able to produce an output within about 50Hz of the nominal 1kHz. If you don't have either an oscilloscope or a digital frequency meter it is possible to use a keyboard musical instrument to set the output frequency since the B5 key (two octaves above middle-C) should produce sound at approximately 988Hz. By comparing the sound from a loudspeaker driven by the test signal and an amplifier from the keyboard instrument you should be able to produce a signal that is very close to 1kHz.

Next month

In next month's *Teach-In 2018* we will be looking at radio frequency (RF) tests and measurements. We will be introducing a selection of RF test instruments and measurement techniques and our practical project will feature a sensitive RF 'sniffer' that can be used to check for radiated signals over a very wide frequency range.

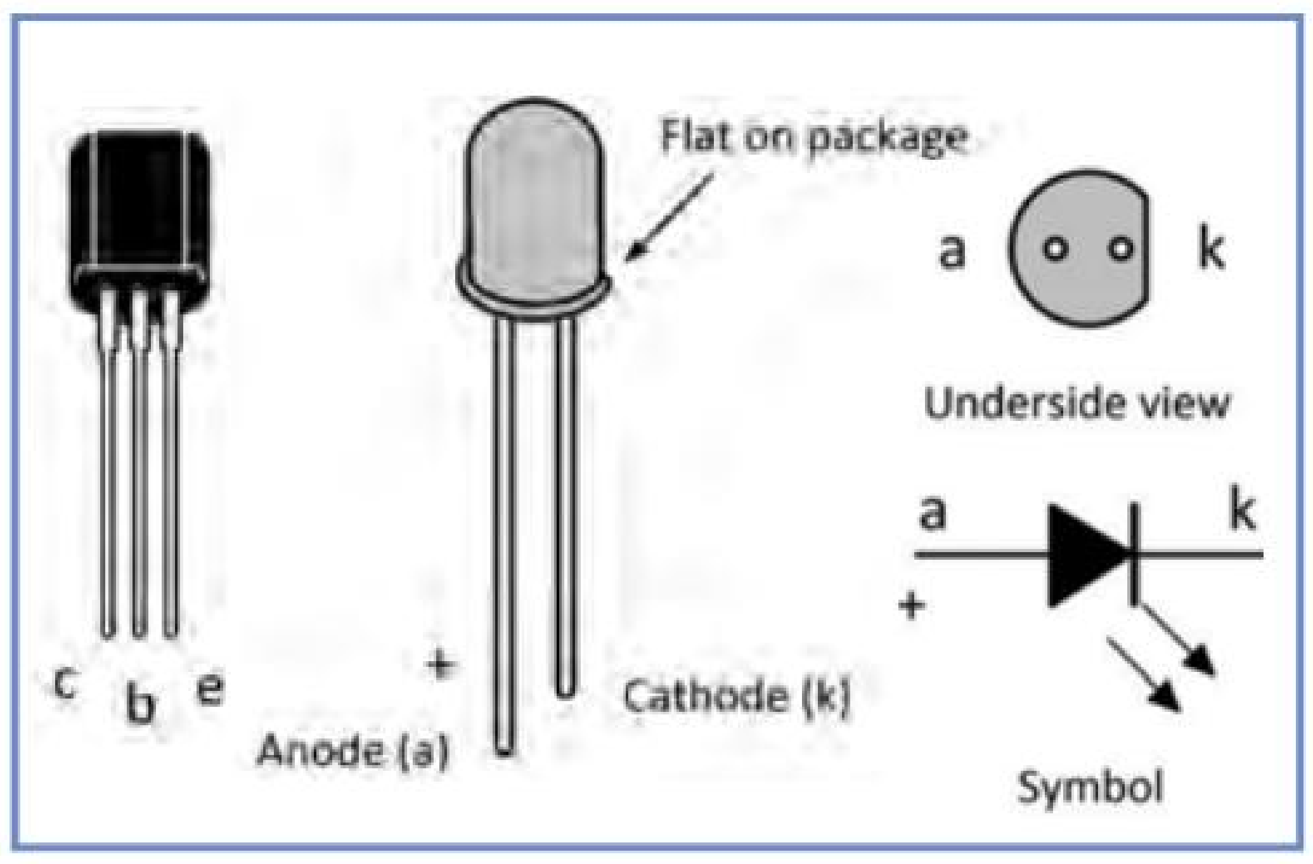


Fig.6.25. (left) Transistor TR1 (BC548) and (middle/right) diode D1 pin connections

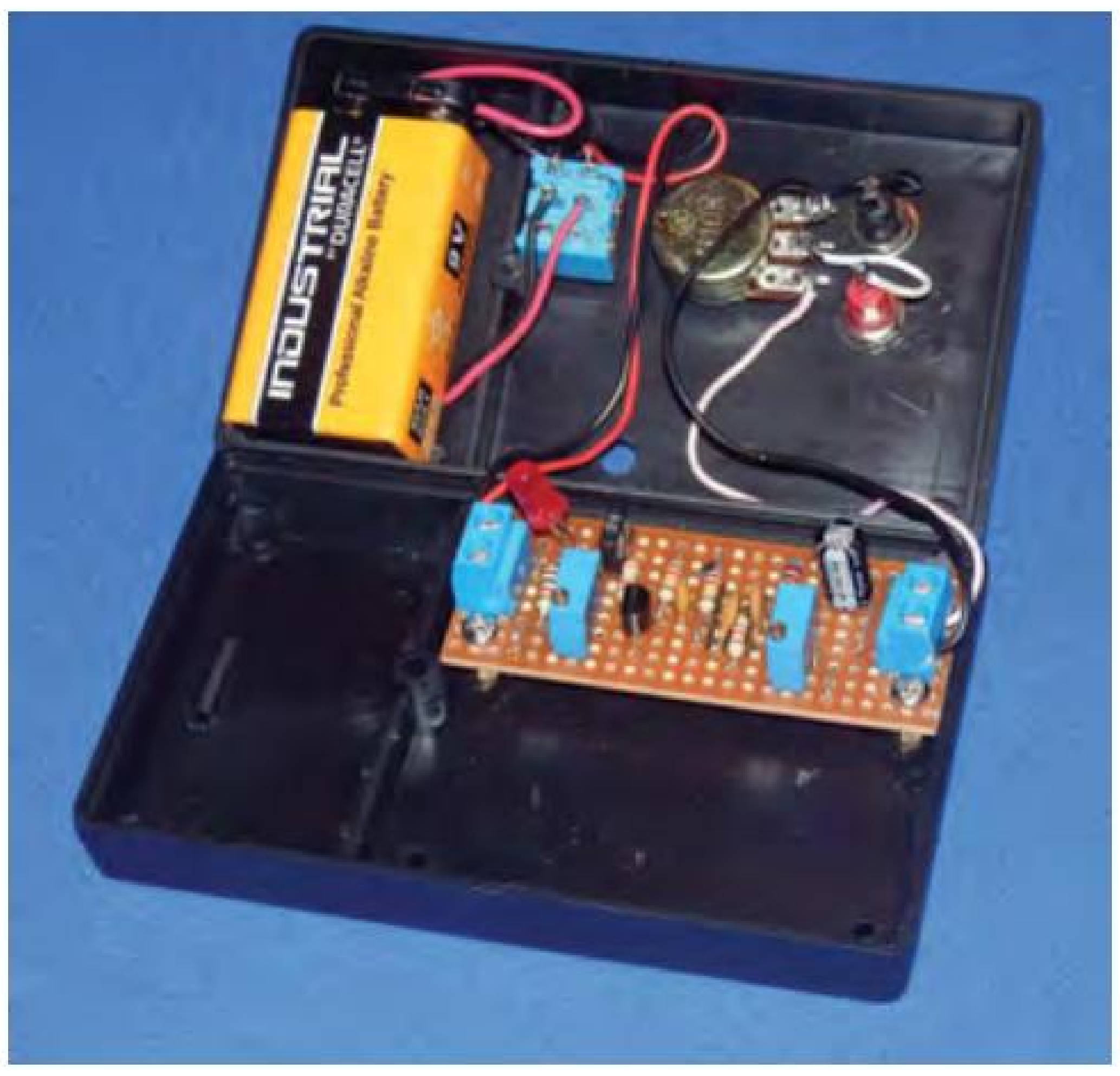


Fig.6.26. Internal wiring of the handy test signal source

EPE
EVERYDAY PRACTICAL ELECTRONICS
ON
LEVEL
OUTPUT

Fig.6.27. External appearance of the test signal source

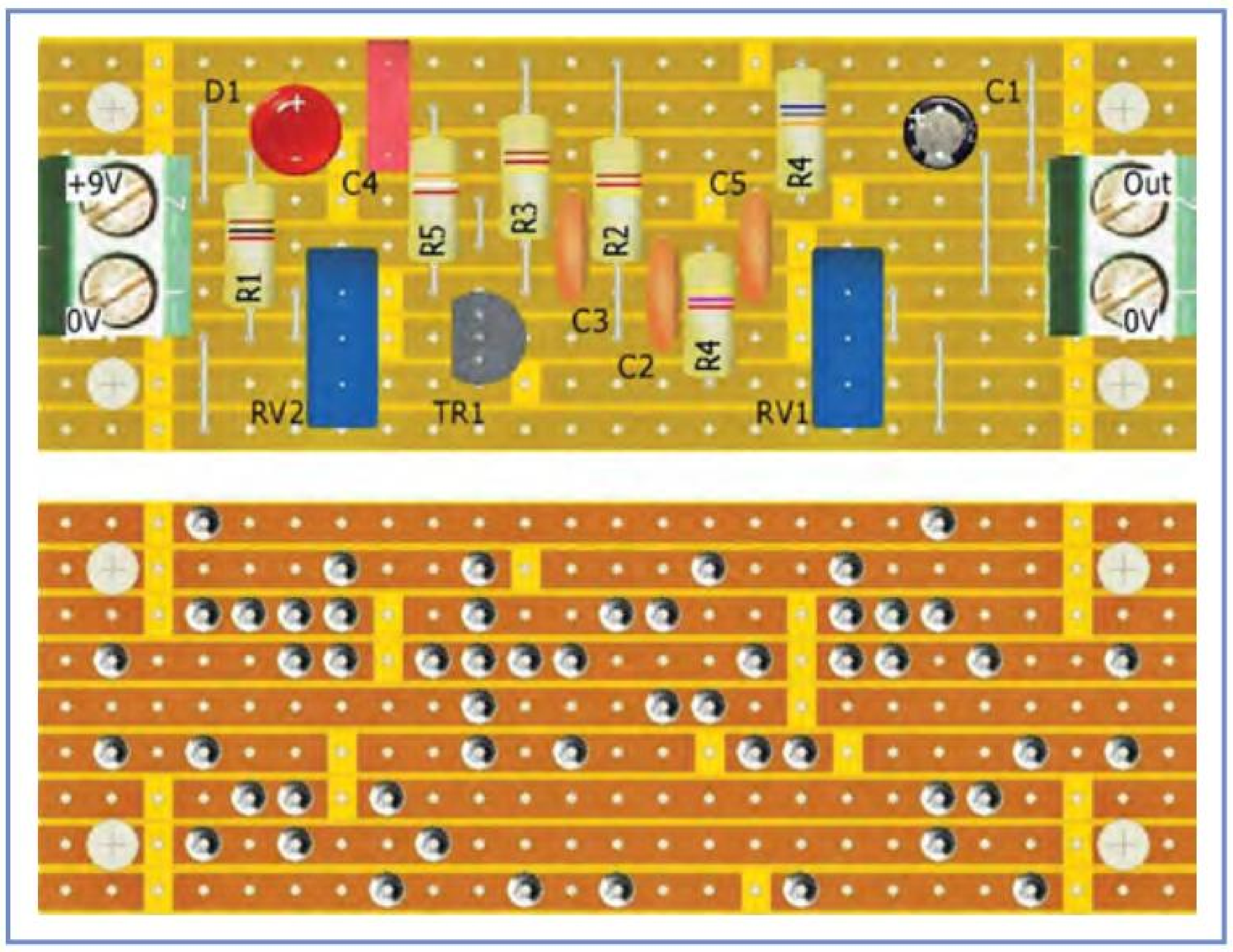


Fig.6.24. (top) Stripboard layout of the handy test signal source and (bottom) underside of the stripboard showing track breaks

Fig.6.28. Using the signal source to test a high-quality audio amplifier at its full rated output power

NET WORK

by Alan Winstanley

Burgers with everything

LAST MONTH, I briefly mentioned the original *Space Spartans* speech-enhanced video game sold in 1982 for the Mattel Intellivision TV console. Another well-loved vintage Mattel TV game was *Burgertime*, and remakes of this old classic arcade game can be found online – a near-perfect Windows version is Magnus Christensen's *Burgertimes* and a small .zip file can be downloaded from: **http://webpages.charter.net/burgertime/btime/burgertimes.htm**

Surfing around the web on a tablet or smartphone, it's curious, and not a little bothersome, to witness how more and more websites are becoming self-contained 'apps' in their own right. Instead of offering visitors simple intuitive navigation, the onset of smaller screens has fuelled demand for compact flyout menus skulking underneath a non-descript 'Menu' button. Bright web developers are rolling navigation aids into those three-lined (or three-dotted) icons, which fly out to offer users a navigational sub-menu intended to ease your way around. Readers will doubtless have seen them in many desktop computer programs (or 'apps' as Microsoft insists we now call them), and many flat, two-dimensional Metro-styled programs are also cursed with these stripey navigation buttons.

In the web design industry the icons themselves are termed 'hamburgers' due to their three-layer appearance. Many websites incorporate so-called 'responsive web design', which detects the platform that the visitor is using (desktop PC, mobile device) and launches a version of the website optimised for the visitor's browser. Thus, small screens now include hamburgers, though desktop browsers such as Firefox and Chrome and many other programs have their own hamburger menu button as well.

According to writer Kelsey Campbell-Dollaghan in an interesting piece written for Gizmodo, the original hamburger flyout icon was a product of Xerox and their Star user interface back in 1981. Some interesting history is presented at: **https://goo.gl/wZAGKR** with a fascinating video explaining Xerox design trends that were extremely advanced for their time and are still with us today. The origin of the simple three-lined design of the hamburger (on the video at 21:15) is put down to the fact that Xerox had very few screen pixels to play with and wanted to show a clear, simple icon. Next time you struggle to navigate with (or even find) a hamburger button on a fancy new website, you can perhaps console yourself with the thought that, for all their cleverness, no-one has improved on a simple design idea dating back to the early 80s.

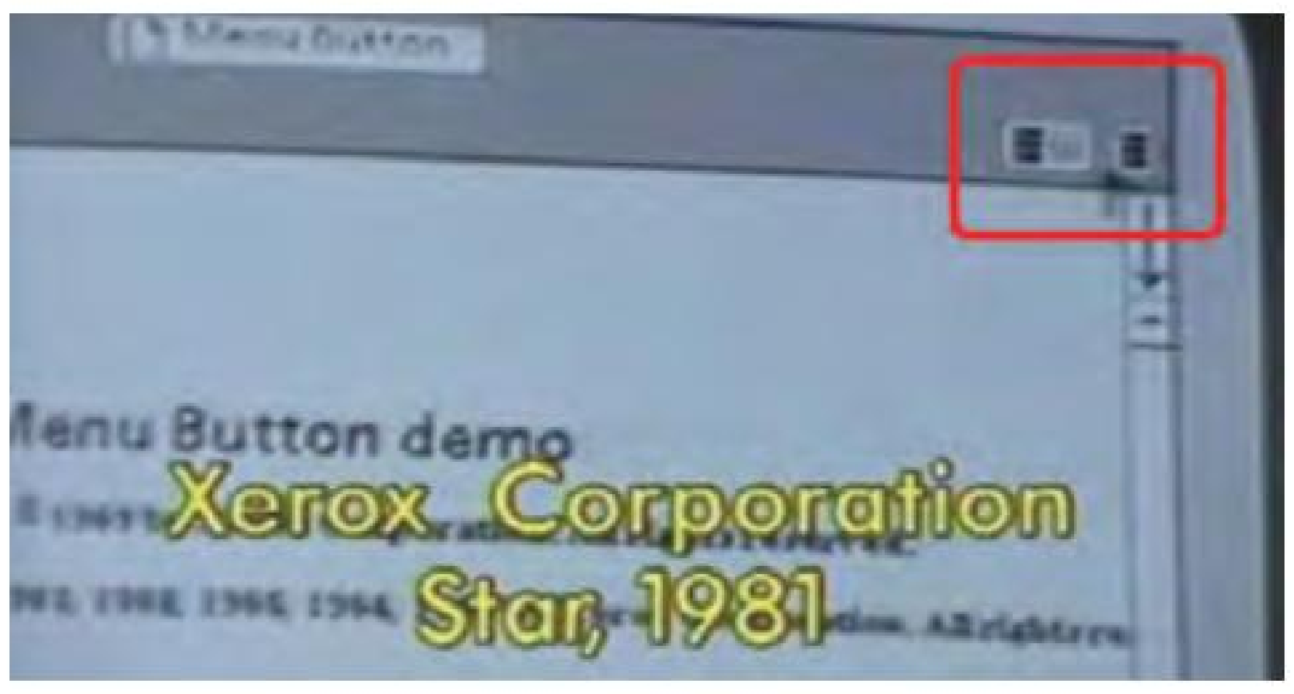
Going Dotty

The author's Amazon Echo Dot – a December Black Friday purchase – continues to behave reasonably well and generally intrigues anyone who sees it. That said, a hard-to-impress cleaning lady sweeping around it with a duster, simply asked if it's one of those 'talking things'.

Its voice recognition is nothing short of remarkable, and Alexa deals with all manner of gruff commands with aplomb – most of the time, anyway. The author did go through a brief phase of cursing dark oaths as the Echo Dot woke up when addressed, but then returned to snooze mode when asked to do something. There followed the usual exasperating ceremony of checking routers, Wi-Fi, IP addresses, rebooting and re-installing before the penny finally dropped: I had been asking 'Alexia' [sic] and the Echo Dot would arouse from her slumbers but then doze off again, ignoring all my commands. Silly me!

The Dot also offers simple on/off control of a bundled TP Link mains socket, although the app has had to be re-installed and set up again at least once. A lack of obvious (and surely basic) functionality detracts from its overall usefulness. Ask Alexa to turn off the socket after half an hour and it cannot: it just turns it off right away. Some timing or countdown commands would enhance its feature set.

As a source of reference knowledge, Amazon's Alexa proves to be somewhat hit and miss. It defines 'soldering' as 'fastening firmly together' (Google Assistant unhelpfully says 'join with solder') and it took several attempts to define 'welding' rather vaguely, but it did define 'printed circuit board' satisfactorily. The Google Chrome browser on a PC can handle voice input from a webcam microphone and is extremely good at handling calculations, thanks to its built-in computational powers. For example, comparing UK/US gasoline prices, when asked 'What is £1.15 per litre in dollars per US gallon?' out pops the answer spoken by Google Assistant of $5.91 per US Gallon. Alexa, however, was thoroughly baffled by this conversion question, though it was a bit more successful with some other searches. There is no denying that Google Assistant is a more powerful and responsive search tool than Alexa is ever likely to be, but then again, Google has had a huge head start on search.



Origins of the 'hamburger' in the Xero Star GUI (graphical user interface) (courtesy Gizmodo)

Light my Fire

An ongoing trade spat between Amazon and Google has brought the news that Google would now block YouTube from appearing directly on Amazon's Echo Show, its new smart device with built-in LCD screen that is ideal for watching streaming movies or video clips. Amazon then removed the YouTube app from its Fire TV devices as well, leaving owners also unable to watch YouTube videos directly on a TV screen. All is not lost, though, because the Firefox web browser is now available on Fire TV, allowing users to surf the web on their HD TV set and access YouTube that way instead.

As previously predicted in this column, there was always the likelihood of major smart speaker vendors slugging it out and wanting to dominate a user's home with their own proprietary systems. Amazon, Google and Apple continue to circle the wagons as they try to make sense of the emerging market for domestic smart speakers, and interesting times are ahead.

That's all for this month's *Net Work*. Write and tell us what you think. You can contact the author at: **alan@epemag.net** or the editor at: **editorial@wimborne.co.uk** and your letter could possibly appear in our *Readout* column.

PIC n' Mix

Mike O'Keeffe

Our periodic column for PIC programming enlightenment

Four-digit, seven-segment LED display – Part 4

THE all-on-one-pin interface solution we explored last month allowed us to add a (calculator) keypad to our original display design, which was first used for a 24-hour clock. Now we're ready to program the hardware to turn it into a calculator.

While the additional modifications to the hardware were minimal, the software will not be as trivial. As we add functionality to the code, it's good to take a step back and consider the current behaviour of the code before changes are made.

The code for the 24-hour clock continually rotates around the four digits. In the background there is a 500ms timer used to time seconds and minutes. The code then updates the display every minute to show the current time. This works well for the clock. However, we must ask ourselves a question – will it still work when we change the fundamental operation from a clock to a calculator?

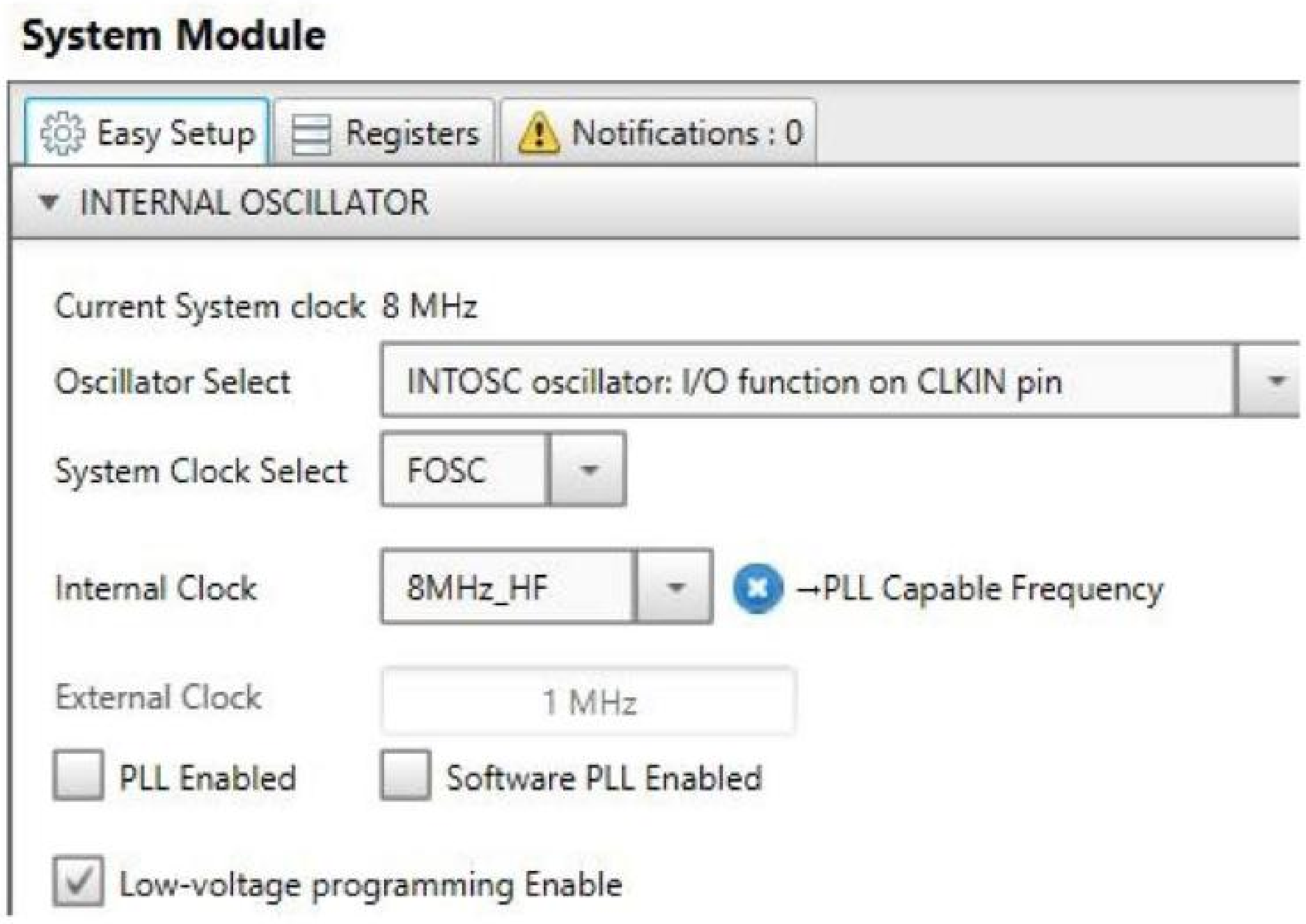
Consider how the calculator should work. Initially, it will display a zero, indicating it is awaiting an input. Pressing any key will commence the equation. The equations for this calculator will be of the type `X ø Y = Z`, where `X` is the first number to be entered, `Y` is the second, `ø` is the mathematical operator and `Z` is the answer. (For our simple calculator, the `ø` can represent addition, subtraction, multiplication or division.) So, we need to capture the first number (`X`), the mathematical operator (`ø`), the second number (`Y`) and the equals sign (=). Once these have been captured, we have enough data to perform the basic equation.

The first number (`X`) in the equation will be complete once an operator key has been pressed, then we can start entering the second number (`Y`) in the equation. Finally, the equals key (=) will be pressed for the answer (`Z`) to be displayed on screen. In the 24-hour clock code, the display is updated every minute. This is far too long to wait for a simple answer, or to update the display with the pressed digit. Therefore, we must update the display much quicker.

Another issue is that we must continually check for a button press using the ADC input on RA0. In the original 24-hour clock code, we manually rotated around the digits, any delay in this would cause one digit to shine brighter than the other or even worse, cause flicker.

Swapping the fundamental behaviour

The answer lies in swapping the behaviour between the 24-hour clock code and the calculator. In the 24-hour clock, the timer is used to count the seconds. The main code then rotates around the digits. The timer uses an interrupt service routine (ISR), which is meant to be very lightweight. In any ISR, the goal is to minimise the interruption to the main code by checking the interrupt flag, perform a very quick operation, reset the flag and return to the main code. If we added the digit rotation into the timer interrupt routine, the seconds would no longer be timed correctly and the clock would have been wrong. However, with the calculator, clock timing isn't important, while performing prompt calculations and ADC captures are. For the calculator, it is better to move our digit rotation into the timer ISR. Now the timer can be set to 1ms for a 1kHz switching frequency. Every time the ISR is called, the digit will swap. This automatic change gives us much more freedom to perform calculations and capture and display the numbers we enter.


System Module

Current System clock 8 MHz

Oscillator Select: INTOSC oscillator: I/O function on CLKIN pin

System Clock Select: FOSC

Internal Clock: 8MHz_HF →PLL Capable Frequency

External Clock: 1 MHz

☐ PLL Enabled ☐ Software PLL Enabled

☑ Low-voltage programming Enable

Fig.1. Changing the internal oscillator in MCC

Moving the digit rotation and display into the ISR adds significant delays inside the ISR. With the clock running at 500kHz and the ISR being called every 1ms, the ISR ends up taking 0.5ms to operate (this was measured during testing using the PICkit3 debugger). If the Timer is set to interrupt every 1ms, then our code only has 0.5ms operation time. The problem here is the main code is constantly interrupted, making it difficult to behave normally. This is a fine example of why ISRs need to be lightweight and fast. With this current clock speed, the calculator will not work properly and updating the display becomes near impossible. Increasing the clock speed to the internal clock at 8MHz improves this operating time to 0.2ms (See Fig.1). However, this is still not ideal and very slow for an ISR.

Entering numbers

It may seem trivial, but pressing the number keys on the keypad and expecting them to appear on the segment display correctly is a little tricky. Consider pressing the key '7' first, then '8' followed by '9'. We would expect the number 789 to appear on the display. In order to display this, we set up a four-digit array in the code and map these to the display's digits. In the array, we have four elements, numbered 0, 1, 2 and 3, where 0 is the first element in the array. In the four-digit seven-segment display, we can map this left-most digit to element 0 in the array and so on until element 3, which is the right-most digit. In our example, '7' would be entered into element 3 first. When we press key '8', the '7' on the display must be shifted into element 2 and the '8' will be stored in element 3. When key '9' is pressed, '7' and '8' will be shifted to the left and '9' will now be stored in element 3. In this example, element 0 does not have a number stored in it. There are two options here, we could display a zero or display nothing (blank LED). Storing a zero in element 0, will display a zero, so we need to choose another digit to represent nothing, which we'll cover further on.

We've now entered a number onto the display. The only problem is this number is split up into an array. We actually have 3 numbers, '7','8' and '9', but what we want for calculations is '789'. We need to convert '7','8' and '9' into a number that our microcontroller can mathematically process. A small function will provide this. The function

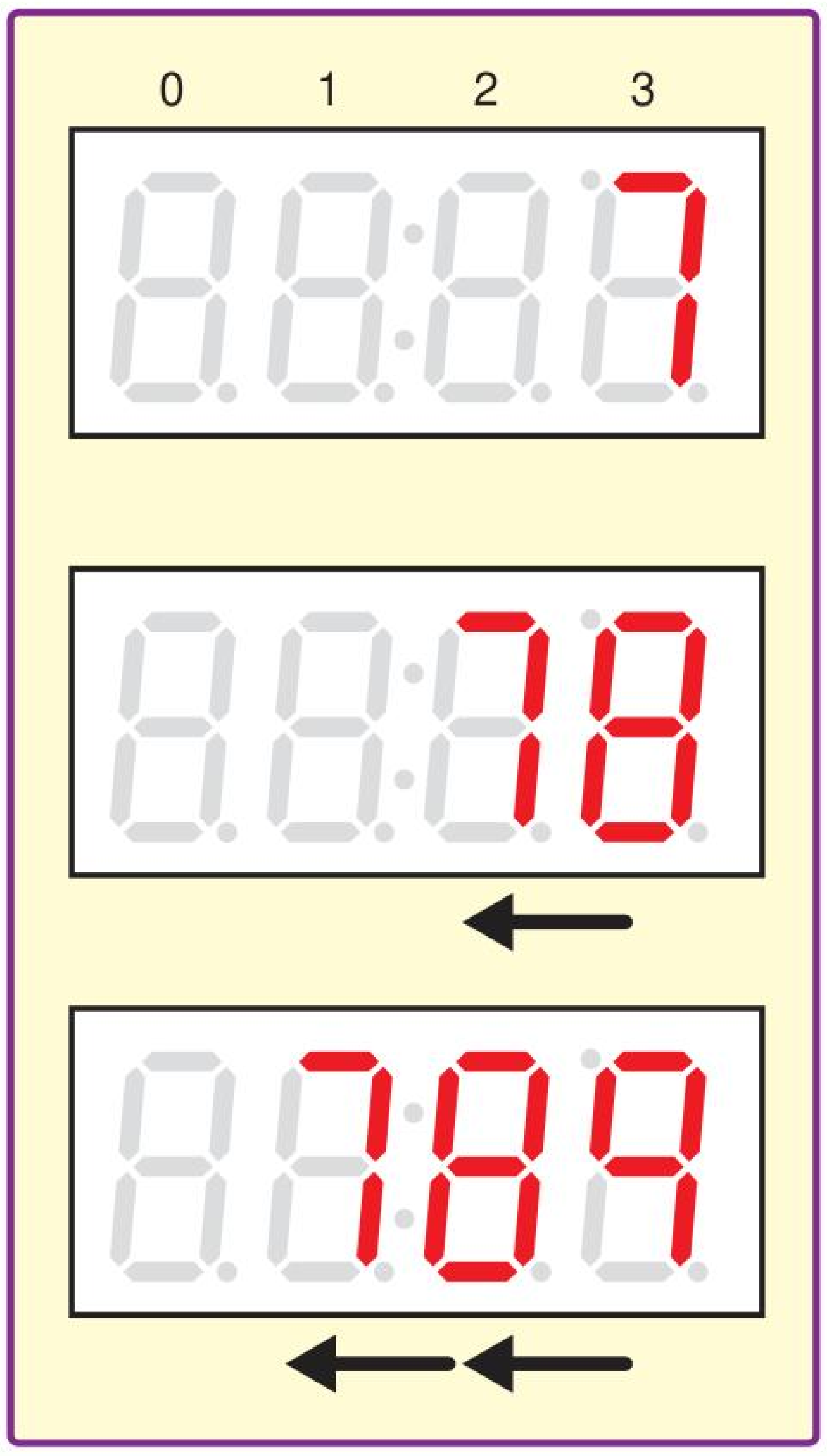


Fig.2. LED display digit shifting

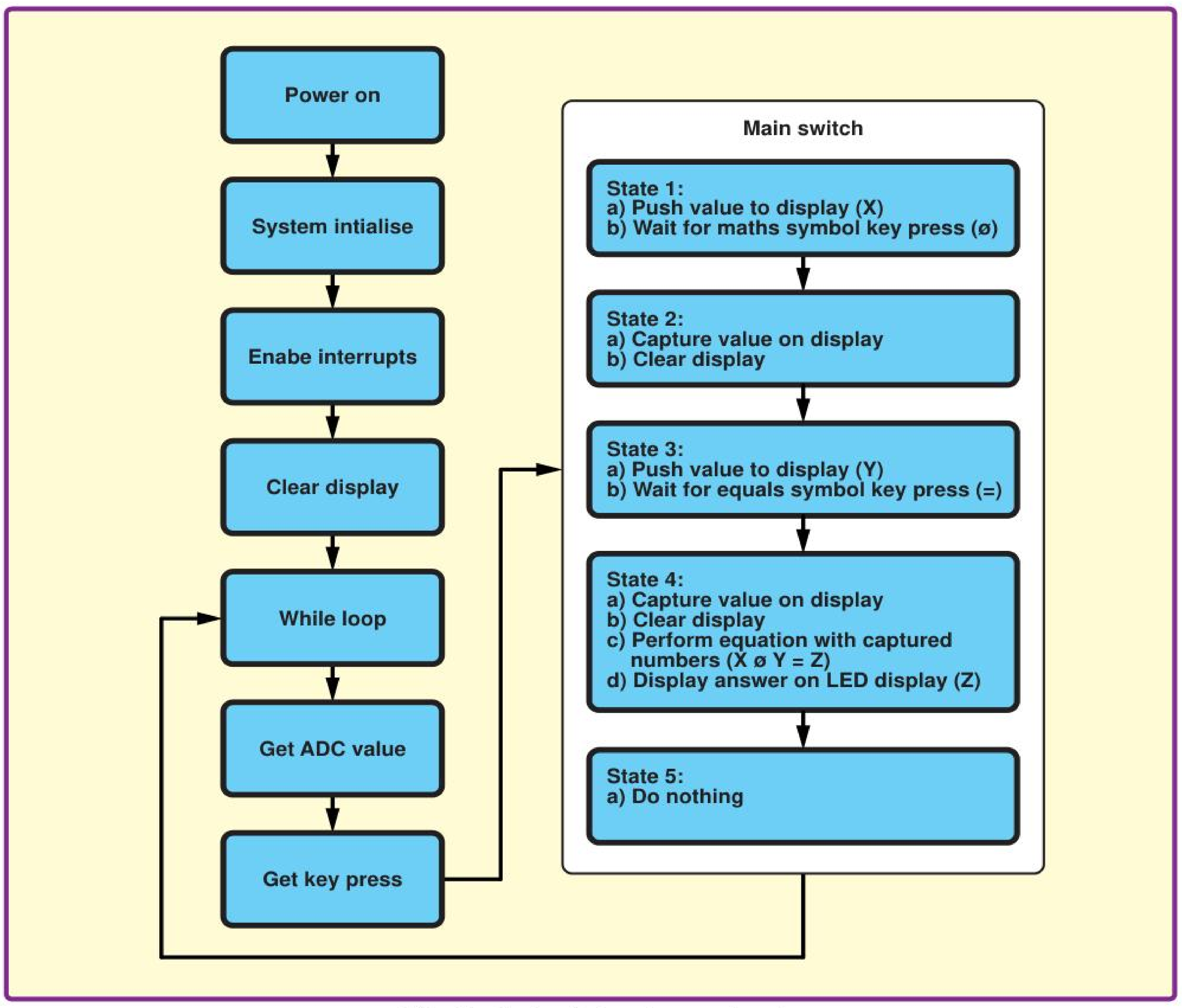

Power on
System intialise
Enabe interrupts
Clear display
While loop
Get ADC value
Get key press
Main switch
State 1:
a) Push value to display (X)
b) Wait for maths symbol key press (ø)
State 2:
a) Capture value on display
b) Clear display
State 3:
a) Push value to display (Y)
b) Wait for equals symbol key press (=)
State 4:
a) Capture value on display
b) Clear display
c) Perform equation with captured numbers (X ø Y = Z)
d) Display answer on LED display (Z)
State 5:
a) Do nothing

Fig.3. Calculator from chart

simply multiplies whatever number (0-9) is in element 0 by 1000. Element 1 will be multiplied by 100, element 2 will be multiplied by 10 and element 4 will be left alone. The results are added to give the result of 789. Now we have our number to work with.

Fig.2 shows the display as each digit is entered, shifting each previous number to the left.

Extra digits

In the 24-hour clock, we only needed the numbers 0-9 to be displayed on the display. This is still the case. However, the issue here is that we are using code from the original program, which is capturing a key press and then displaying it on the display. The normal digits will be mapped but the mathematical operator key presses now need to be added for the calculator to be able to work.

The original code uses a switch statement to swap between the various numbers to be displayed. The function `displayNumber()` is used to display the numbers on the display. In reality, it is used to convert key presses to exact pins connected to the segments to be controlled. To build upon that, we need to be able to capture when the multiply, divide, subtract, add, equals and clear buttons are pressed. We also want to be able to show a blank digit (ie, no LEDs lit) as it is easier to read a number on the display with no leading zeros. For improved calculator functionality it's a good idea to try and display an error message if a calculation goes wrong. This can be done by displayed the letters 'Err' as a shorthand for error. This will be called when the resultant number is larger than the four-digit maximum or when there's a divide-by-zero scenario.

```
0xA  => Equals
0xB  => CLR
0xC  => Multiply
0xD  => Divide
0xE  => Subtract
0xF  => Add
0x10  => E
0x11  => r
0x12  => Nothing
```

The above maps out the new cases to be added to the function `displayNumber()`, where 0xA now represents equals. In the code, the case switch will perform some operation when this is seen, as with all the other values.

The code

There are significant changes to the code in order to convert it to a calculator. We need to move the digit display and rotate function into the Timer0 ISR. We will also need to completely change what's happening in the main code in order to control the behaviour of the project. We will be using a state machine (see Fig.3.). Space considerations mean it is not feasible to look at every change made to the code, but we will look at some of the key features.

```
void TMR0_ISR(void) {
    INTCONbits.TMR0IF = 0;
    TMR0 = timer0ReloadVal;

displayDigit(currentdigit, digits[currentdigit]);

    if(currentdigit >= 3) {
        currentdigit = 0;
    } else {
        currentdigit++;
    }

    if(TMR0_InterruptHandler) {
        TMR0_InterruptHandler();
    }
}
```

Instead of incrementing the seconds, the Timer0 ISR now calls the function `displayDigit()`. The variable `currentdigit` is incremented in the `if` statement that follows, making sure it circulates from 0 to 3 and restarts at 0 again.

```
clrDisplay();
calcstate = 1;
```

Starting off with the main code, we're going to first use the `clrDisplay()` function to turn off digits 0-2 and place a waiting 0 on digit 3. This will be our initial starting point. Since we will be using a state machine, we need to start it off by setting it to the first position using the `calcstate` variable.

```
while (1) {
    KeypadVal = ADC_GetConversion(0);
    disVal = getKeyPress(KeypadVal);
```

The `while` loop starts with an ADC capture using the MCC `ADC_GetConversion()` function. The value is then stored in `KeypadVal`. The next function called is `getKeyPress()`, which takes the value in `KeypadVal` and tries to evaluate which button has been pressed. These two function calls are outside the main switch, meaning they will always be called. `getKeyPress()` will convert any key press to a specific value (all except one key, which is the CLR key). When the CLR key is pressed, a soft reset occurs using the assembly soft reset command `asm ("reset");`. A software (or **soft**) reset is one where the code jumps to instruction zero. (A **hard** reset, is where the power is cycled on and off. A soft reset will often start everything from fresh, but it may not always work, especially if poorly designed code writes over parts of memory that it shouldn't, thereby corrupting the memory space.)

```
switch(calcstate) {
case 1:
    if(mathSign > 0) {
        calcstate = 2;
        break;
    }
    if(KeypadVal < maxADC) {
pushToDisplay(disVal);
    }
    break;
```

Starting with Case 1 in the switch statement, we want to display the numbers captured and converted in the previous functions. The value to be displayed is stored in the variable `disVal`. First, we check to see if the value `mathSign` has been assigned a value. `mathSign` is initialised as zero in the `functions.c` file. When a mathematical operator key has been pressed, `mathSign` will be assigned a value based on the key pressed. At this point, the next state will be selected. Before that, we check the `KeypadVal` is less than `maxADC`. This verifies a valid number key has been pressed. `maxADC` represents the maximum ADC input for a valid key press. Then the value is pushed to the display using the `pushToDisplay()` function. This function will not be discussed here, it simply left shifts any current digits and stores the new value in the right-most digit. It will only allow four values, maximum. Any numbers pressed after that will be ignored.

```
case 2:
    getDisplayNumber();
    clrDisplay();
    __delay_ms(100);
    calcstate = 3;
    break;
```

In Case 2, we want to grab the number on the display. This is currently stored in a 4-byte array called `digits[]`. The `getDisplayNumber()` function takes the separate numbers in the `digits[]` array and converts them into a single number. For example, 7, 8 and 9 would be combined to get 789. The display is then cleared using `clrDisplay()` – again, a small wait and the state machine moves onto the next state.

```
case 3:
    if(disVal == 0xA) {
        calcstate = 4;
        break;
    }
    if(KeypadVal < maxADC) {

pushToDisplay(disVal);
    }
    break;
```

In Case 3, we want to capture the second number and display it on the LED display. This is similar to Case 1, except we're looking specifically for the equals key to be pressed (which is represented by 0xA as mentioned earlier).

```
case 4:
    getDisplayNumber();
    clrDisplay();
    mathAnswer = performMath();
    convertDisplayNum(mathAnswer);
    calcstate = 5;
    break;
```

Case 4, we want to capture the second number on the display. We clear the display using the `clrDisplay()` function again. Next up we have a function caller `performMath()`, which takes the first number entered, the captured math operator and the second number entered and evaluate the answer, which is then stored in the variable `mathAnswer`. This number must now be converted into a format that can be displayed on the LED display. `convertDisplayNum()` is the function that converts the number into the 4-byte array `digits[]`. To finish, we move onto the next state using the `calcstate` variable again.

```
case 5:
    // Do nothing here
    break;
```

This is an important state in the state machine. Here we enter a state, from which we will not easily exit. At this point, the result will be displayed on the LED display. The only way to get out of this is to press the CLR key, which will reset the PIC and the calculator.

```
default:
    calcstate = 1;
    break;
}
}
```

It's not always necessary to add the default case in a switch statement, but it is good practice. If for some weird reason the variable `calcstate` contains a value other than 1-5, then the default case will reset this variable back to 1, resetting the process again.

There's a few interesting points to see in the software. One of the key points from above is keeping the ISR as lightweight and as fast as possible. Although adding more functionality to the ISR can be good providing the risks and delays are understood. It's also interesting to note the changes to the hardware were minimal, while the software changes were rather extensive. There's very little code that could be re-used.

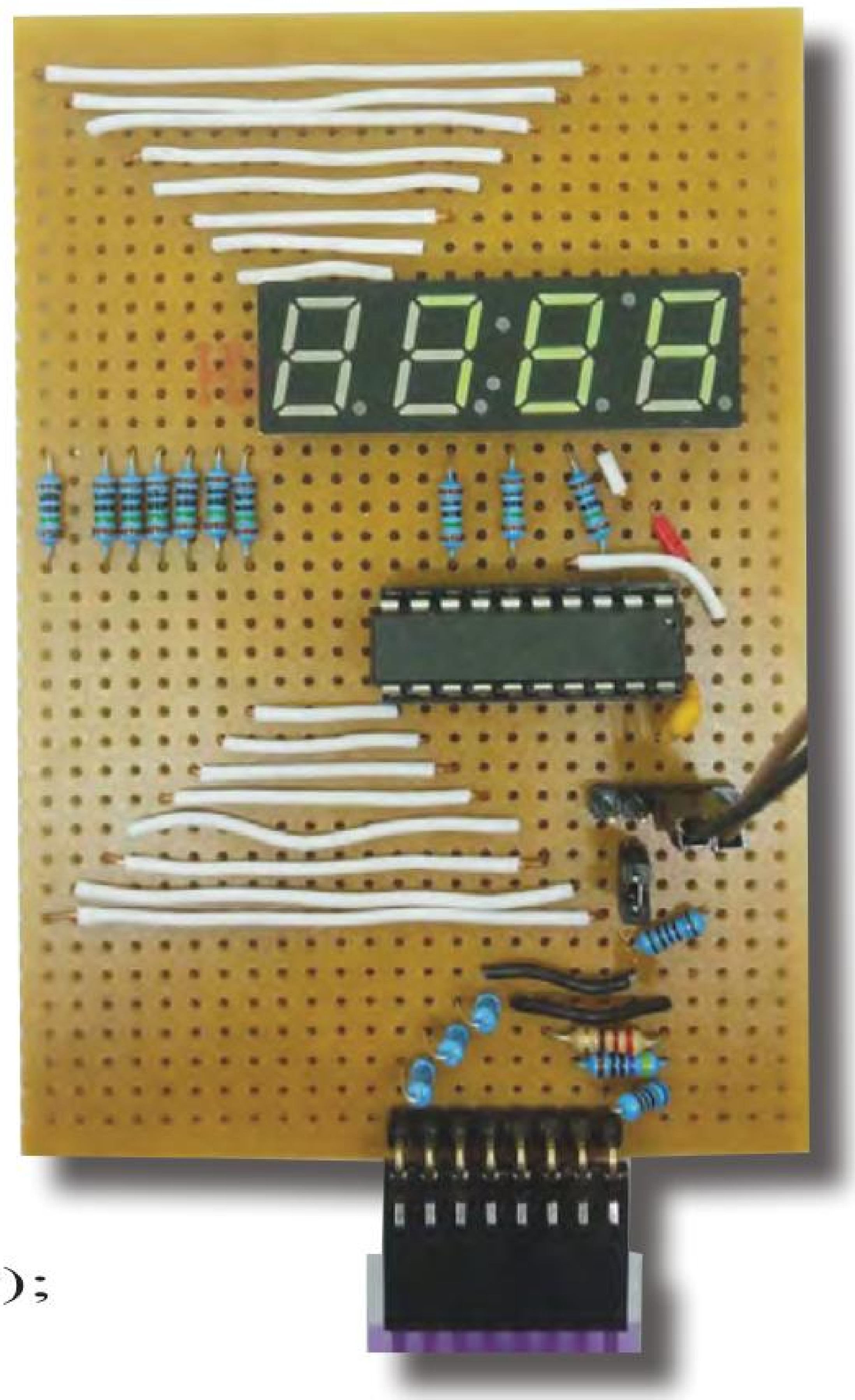


Fig.4. Fully working calculator showing 789 as the result

Last but not least, the display is noticeably brighter. Having the display and rotation in the ISR ensures that each digit is given equal power-on time.

Next month

I'm taking a small sabbatical for my greatest adventure yet – the birth of my newborn twins: Chris and Ethan. While I'm away, the original and highly esteemed *PIC 'n Mix* columnist Mike Hibbett will be making a short return to fill in for me. He has some exciting projects in store for you. All I'll say is that it has something to do with 'FFT'. I look forward to seeing him back in action and I will see you all again upon my return.

Not all of Mike's technology tinkering and discussions make it to print.

You can follow the rest of it on Twitter at *@MikePOKeeffe*,

on the *EPE Chat Zone* or *EEWeb's* forums as *'mikepokeeffe'*

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CIRCUIT SURGERY

REGULAR CLINIC

BY IAN BELL

Power supply switching

LAST MONTH, we considered some issues related to the provision of multiple supply voltages. This is a very common requirement in electronic systems where different chips and subsystems, and often individual chips, need different power supply voltages. That article was in response to some discussion on the provision of multiple power supply voltages, power supply switching and regulator circuits, started by user ***Tuurbo46*** on the *EPE Chat Zone*. The discussion continued (under username ***Rocket Ron***) on the new *EPE* forum hosted by EEWeb (**www.eeweb.com/forum/tags/epe-magazine**).

One of the specific issues from ***Tuurbo46***/***Rocket Ron*** posts concerned the possible use of a potential divider to obtain a lower supply voltage from a higher one. We looked at the fundamentals of potential dividers and demonstrated that potential dividers were very inefficient (wasteful of power) used in this way if reasonable load regulation was required, and are thus generally unsuitable for providing power supply voltages. The exception might be for a sub-circuit with extremely low power requirements. We also looked briefly at possible configurations of two regulators used to provide two different supply voltages.

On the EEWeb forum the discussion moved on to the issue of switching power supplies on and off using a microcontroller. ***Rocket Ron*** wrote: 'I am now going to use three LM317s to set separate voltages, 5V, 2.5V and 1.25V to supply 100mA. I want to switch these three separate voltages ON/OFF with three micro I/O pins...'.

Later, he added: '...I have decided to use a PFET in series with VIN on the LM317 (default setting off), and then an NPN on the gate connected to GND. The I/O pin is connected to the base of the transistor. When voltage is required, the I/O pin powers the base on the transistor. Is this a better method of doing it?'.

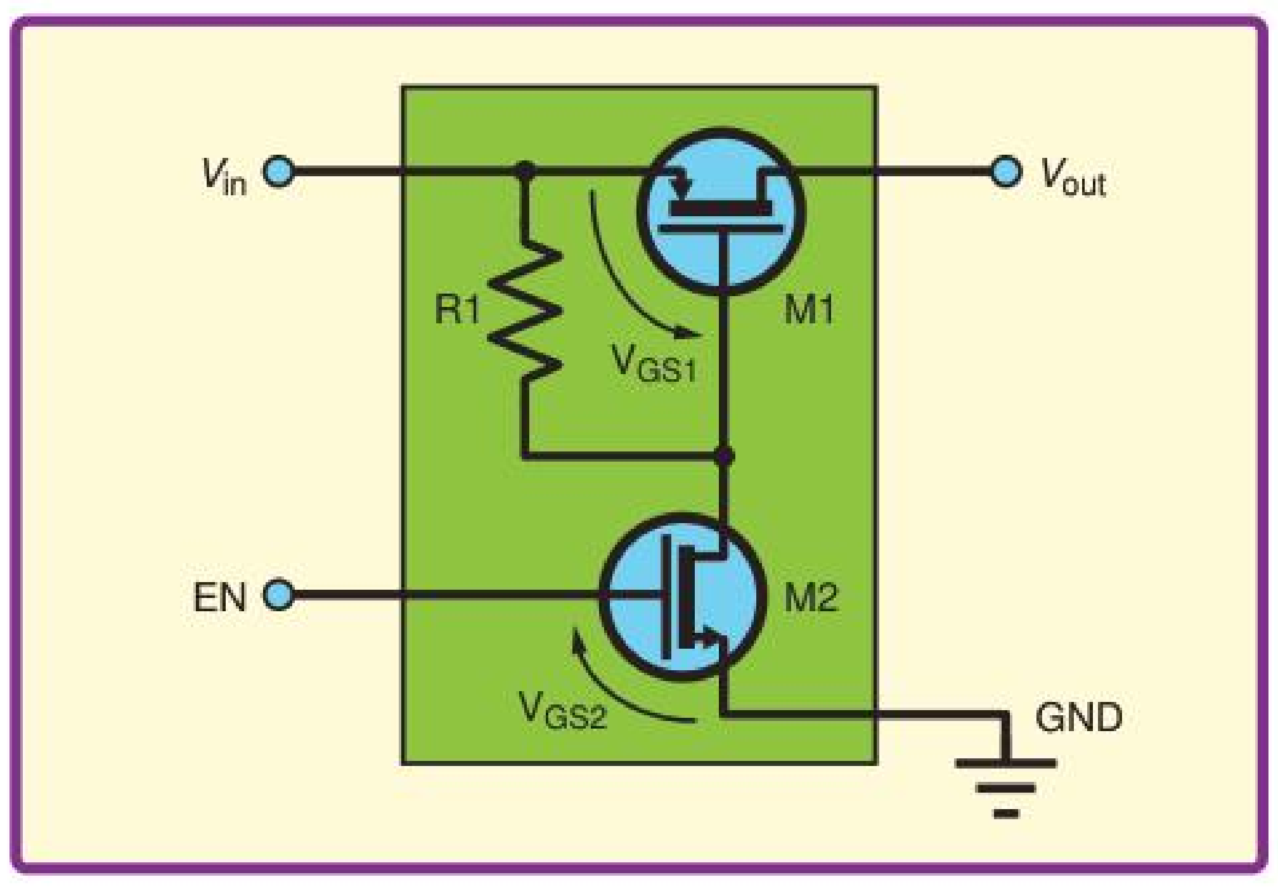

Fig.1. Basic load switch circuit.

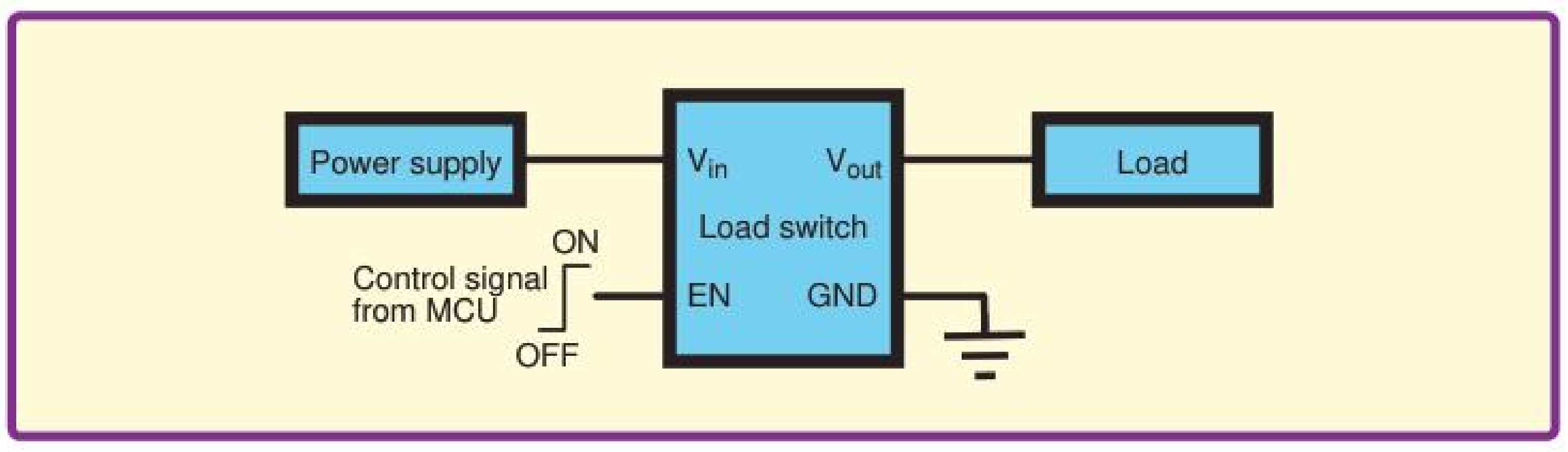

Fig.2. Connection of load switch from Fig.1 in a circuit

Therefore, continuing with the power supply theme from last month, we will now look at the various options for switching power supplies on and off as part of system operation.

The switching of power supplies, like the need for multiple supplies, is a common requirement in electronic systems. Multiple power supplies often have to be switched on and off in a particular order to ensure correct operation, or even prevent damage to a system – this is known as power supply sequencing. It is also a common requirement to be able to power-off some subsystems when they are not in use to reduce power consumption – known as power supply distribution.

PMOS load switches

Circuits for power supply switching are typically called load switches or power switches. You can also implement a load switch using transistors, as suggested by ***Rocket Ron***. A typical basic load switch circuit is shown in Fig.1, and is similar to the circuit described by ***Rocket Ron*** except that both transistors are MOSFETs. EN is the enable input, which may be connected to a microcontroller output (as in ***Rocket Ron's*** application). The connection of the load switch in a circuit is shown in Fig.2.

In the circuit in Fig.1, M1 is a PMOS transistor, which acts as the main switch. M2 is an NMOS transistor used to switch M1 via the EN input. A high input (logic 1 output from a microcontroller general purpose I/O pin) is required at EN to switch the load on (connect V_{in} to V_{out}). With 0V applied (GPIO logic 0) to EN, the load will be off (V_{in} disconnected from V_{out}).

To consider the circuit in more detail recall that a MOSFET transistor will switch on when its gate-source voltage (V_{GS}) is greater than its threshold voltage (V_{TH}). A PMOS transistor requires a negative V_{GS} to turn on (gate negative with respect to source) whereas an NMOS transistor requires a positive V_{GS} (gate positive with respect to source). In the circuit in Fig.1, with a high input on EN, M2 turns on and pulls the gate of M1 low (close to 0V) so that, assuming V_{in} is larger than the threshold voltage of M1, M1 will turn on, connecting V_{in} to V_{out}. When EN is at 0V, M2 will be off (effectively open circuit), during which time resistor R1 ensures that the gate voltage of M1 is pulled up to V_{in} – so that the gate-source voltage of M1 is close to zero and M1 is off. In some situations it may be necessary to include a pull-down resistor between the gate of M2 and ground to make sure M2 is always off when not actively driven.

It may be possible to use a simpler switch circuit – just a PMOS transistor (as in Fig.3) with its gate connected directly to the microcontroller. The connection of the load switch is as in Fig.2 except that the logic of the on and off control on EN input is inverted. Although this circuit uses fewer components it has a number of disadvantages. One problem is that the switch will be on with a 0V output from the microcontroller. This means the switch is likely to

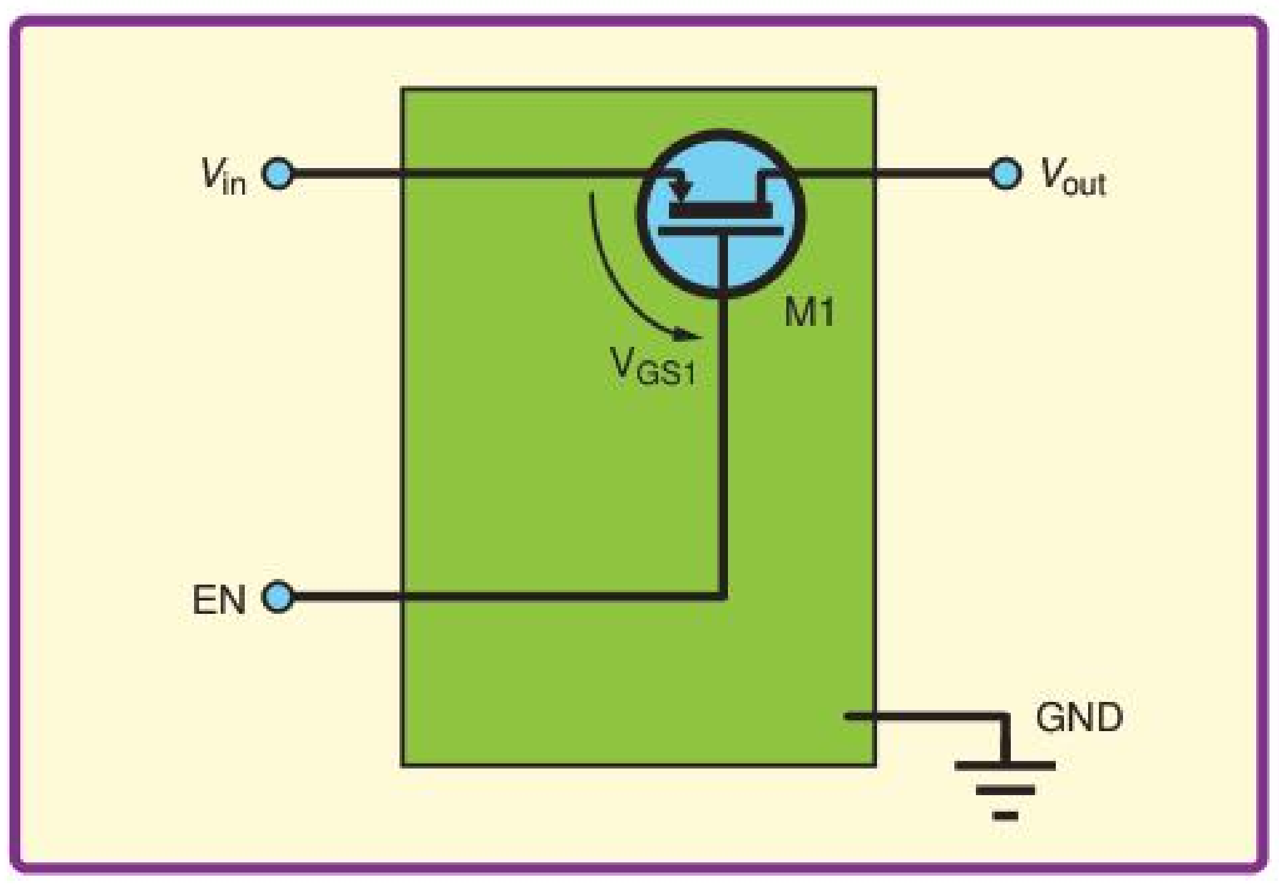

Fig.3. A single PMOS transistor is potentially usable as a load switch, but this circuit has some disadvantages.

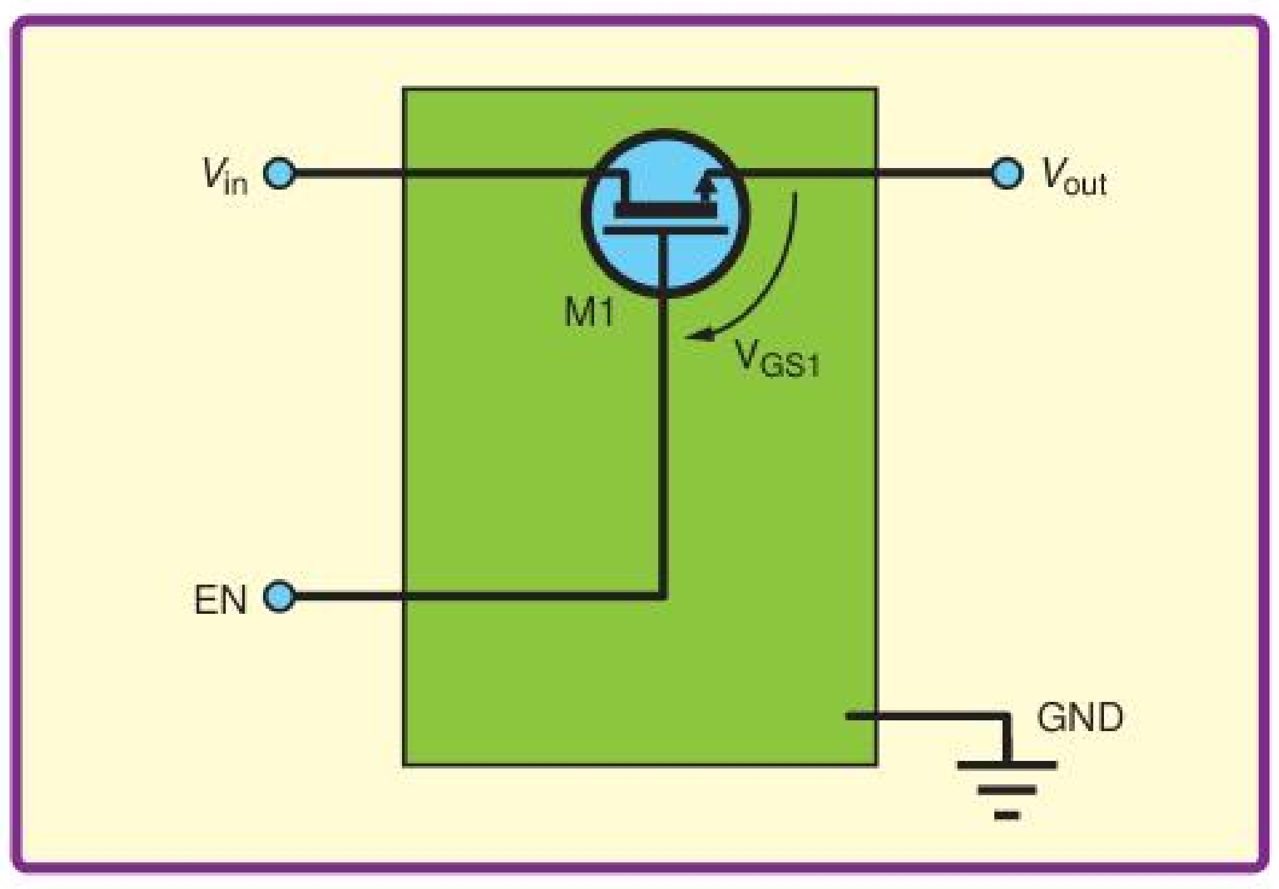

Fig.4. A single NMOS transistor used as a load switch. This circuit can be used if the logic high voltage applied to EN is sufficiently more positive than V_{IN}.

be on at system power-up, before the microcontroller initialises and the code has reached a point where the switch can be controlled. Also, putting the microcontroller into power-down or sleep modes will be likely to turn the switch on, which in many cases would not be what is required.

Another issue with using a single PMOS transistor relates to the voltages that can be switched. For the transistor to be off its gate voltage must be greater than $V_{IN} - V_{TH}$ so that V_{GS} is less than V_{TH}. This restricts the maximum V_{IN} voltage that can be handled to be $V_{OH} + V_{TH}$, where V_{OH} is the logic 1 (high) output voltage from the microcontroller.

NMOS switches

A single NMOS transistor can also be used as a load switch, as shown in Fig.4. The connection and logic direction are as in Fig.2, but because the NMOS transistor requires a positive V_{GS} to switch on, V_{IN} must be less than $V_{OH} - V_{TH}$. This could be feasible in some situations, such as a 5V microcontroller controlling a 3.3V supply, or a 3.3V microcontroller switching a 1.8V supply. However, there are many situations where a single NMOS transistor cannot be used.

In situations where the control voltage is not larger than V_{IN}, an NMOS transistor can still be used with the help of some additional circuitry to produce a higher voltage than V_{IN}, which is used to control the gate of M1 – see Fig.5. The higher voltage is produced by a circuit called a charge pump, a form of switch-mode DC-to-DC converter which only uses capacitors, rather than the inductors commonly found in full switch-mode power supplies. Use of capacitors facilitates on-chip implementation, although discrete-component charge pumps can also be built.

The basic way in which a charge pump works is to charge a capacitor to V_{IN} and then use MOSFET switches to rearrange the capacitor connections so that the negative end is connected to V_{IN}. This produces a voltage of $2V_{IN}$ (across V_{IN} and the capacitor in series), which can be used to charge another capacitor to $2V_{IN}$. The switching process is repeated under control of a clock signal to keep 'pumping up' the output capacitor to the higher voltage. Diodes are used to ensure the charging currents flow as required. Using multiple stages, voltages of several times V_{IN} can be achieved.

So far, we have introduced some basic load-switching circuits, and with the exception of the charge pump in Fig.5, these require a very small number of components and are easily implemented with discrete MOSFETs. However, a variety of load switch integrated circuits are available from a number of manufacturers. The availability of ICs indicates a commercial driver for their existence, in this case it is the common need for load switching combined with miniaturisation – load switch ICs take up less board space than circuits built with discrete components. This is important if you are trying to fit everything in a small package such as a mobile phone or mini tablet. Miniaturisation is generally of less concern to amateur designers, but IC load switches often provide additional features or performance enhancements over the basic two-transistor circuit, so are worth considering.

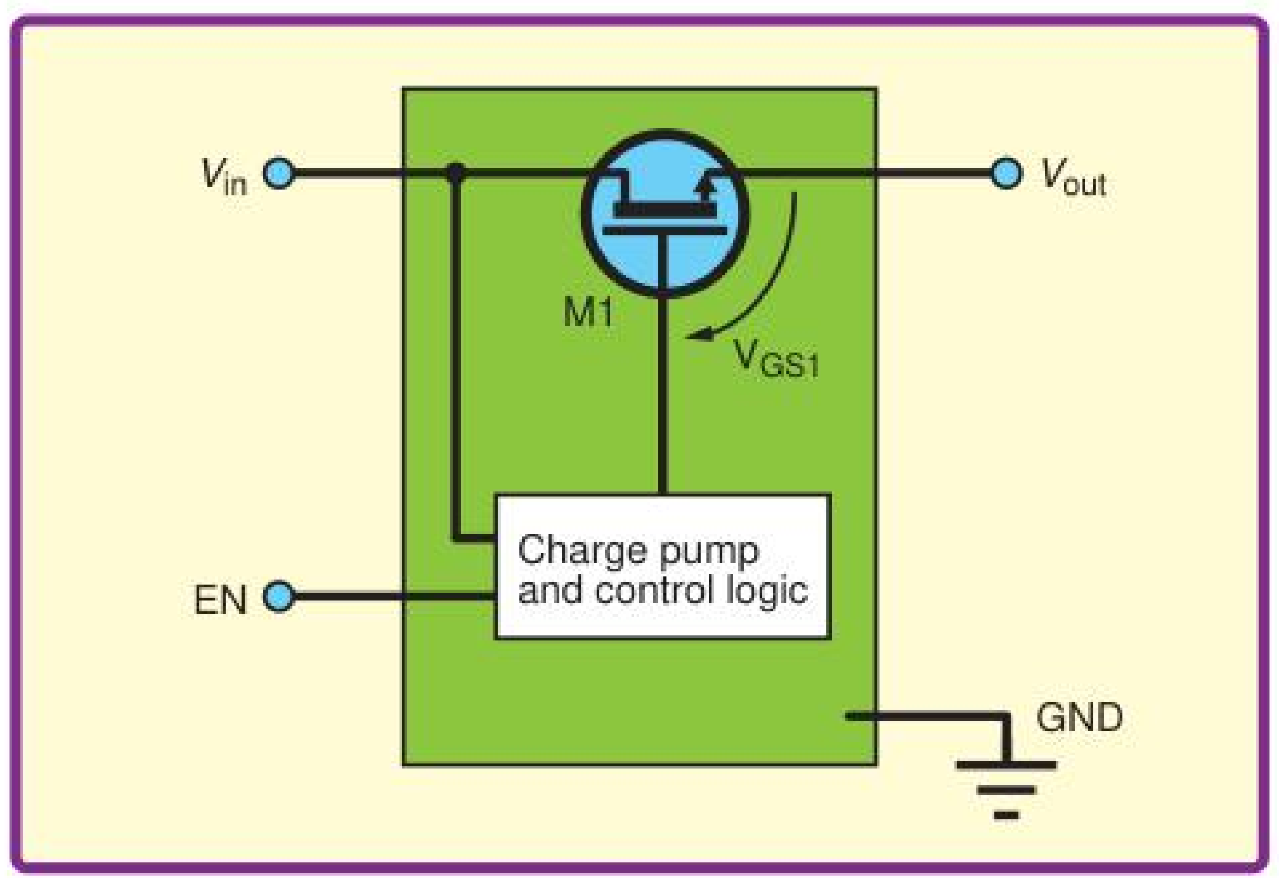

Fig.5. NMOS transistor used as a load switch with a charge pump supplying the required gate voltage.

Characteristics

There are a number of characteristics and issues that may need to be considered when selecting a load switch IC or designing a load switch circuit. Some basic characteristics are detailed below.

Input voltage – the range of voltages, which can be switched. The power supply voltage being switched must be within this range. For discrete designs this will relate to the maximum voltage ratings of the transistors used. For load-switch ICs, the input range will be given on the data sheet. Furthermore, load-switch ICs may also have another connection (sometimes called the bias voltage), which is a supply for the internal circuits. Again, the datasheet will specify the requirements for this voltage.

Maximum continuous current – the maximum current the switch can handle. The maximum continuous current taken from the supply by the circuit being switched must be less than this. This will specified for a load-switch IC. For a discrete MOSFET, the maximum continuous drain-source current must be suitable.

On resistance – the resistance of the load switch from input to output in the on state. This is basically the on resistance of the M1 MOSFET in Fig.1, 3, 4 and 5. The on resistance has a significant effect on the power dissipation of the load switch when the load is active – a low on resistance is required to keep dissipation down. It is worth noting that although M1 (in the circuits shown) is on when V_{GS} is just greater than the threshold voltage, a sufficiently low on resistance may require a significantly larger gate-source voltage than V_{TH}. This could be an issue in some uses of the circuits in Fig.1, 3 and 4, depending on V_{IN} and/or the microcontroller logic voltage. The charge-pump circuit in Fig.5 can overcome the limitations of the low circuit voltages to apply a larger V_{GS} to achieve low on resistance.

Leakage and quiescent currents – the load switch will exhibit some leakage current from V_{IN} when the power supply is on and the load switch is off. For IC load switches there will be some quiescent current in addition to the load current required to power internal circuitry, even if no load is connected. For the circuit in Fig.1, when the switch is on there will be current flowing through R1 (via M2) which will be in addition to the load current, reducing the efficiency of the circuit.

There are also a number of circuit behaviour/performance issues, which may need to be taken into account and which may not be obvious to anyone who has not used load switches before. Of particular importance are: the effect of inrush current, the behaviour of the load after switch-off, and the need for reverse-current protection in some applications.

Inrush current and supply dip

Directly after a circuit is switched on it may briefly take a much higher current than during normal operation. Typically this is due to capacitance across the supply lines changing up and is referred to as 'inrush current'. For a simple system, where all the circuitry powers up together this may not be a problem – the system can be stopped from trying to do anything with a power-on reset which lasts longer than time taken for the supplies to settle. However, when a system is already powered up and a new subsystem is switched on using a load switch, the power supply will experience the inrush current and may not be able to sustain its output voltage due to the sudden heavy demand. The supply voltage will dip momentarily, which may disrupt the operation of other subsystems that

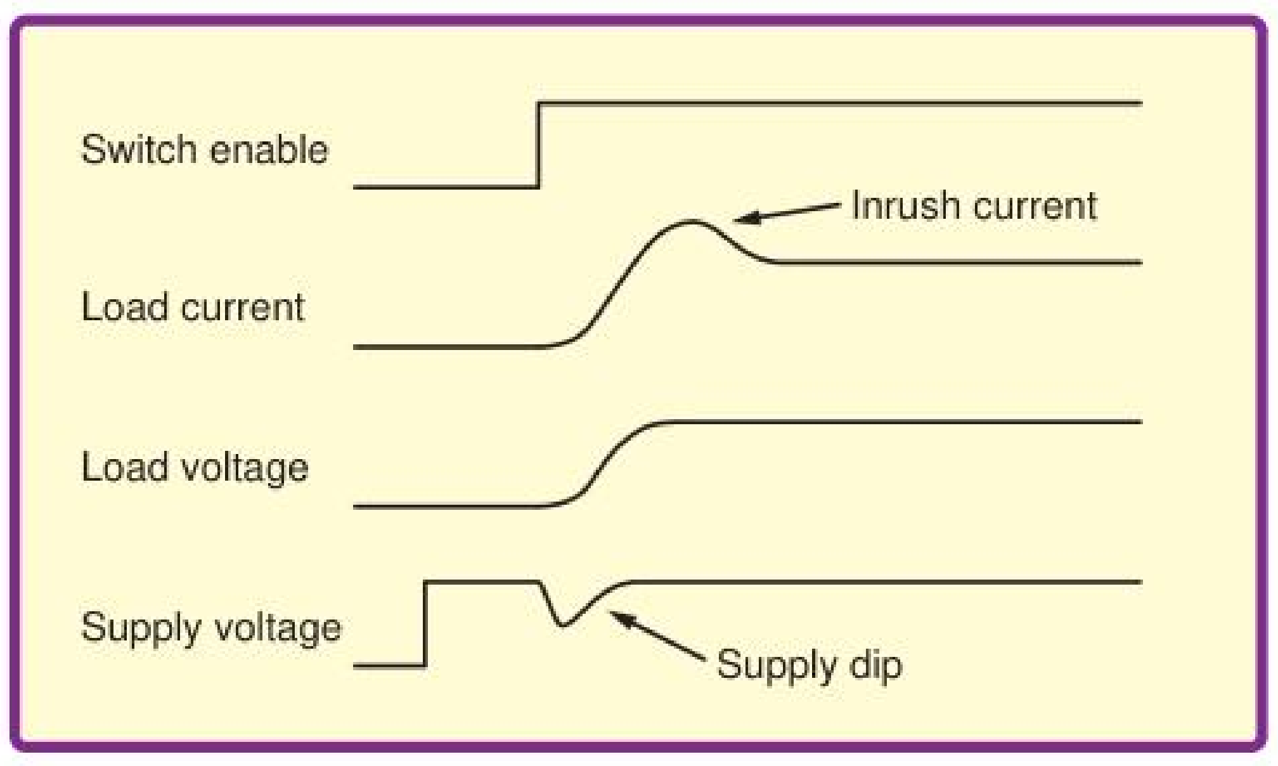

Fig.6. Inrush current and supply dip waveforms.

are already powered up. Waveforms illustrating this situation are shown in Fig.6.

To prevent load switches from disrupting system power supplies they can be designed to limit the rate of increase of their output voltage (known as the 'slew rate'). This is a typical feature of IC load switches, which may allow the slew rate to be controlled via an external capacitor or resistor. In discrete circuits, the output voltage slew rate can be reduced by putting a resistor in series with the MOSFET gate. Fig.7 shows the circuit from Fig.1 modified in this way.

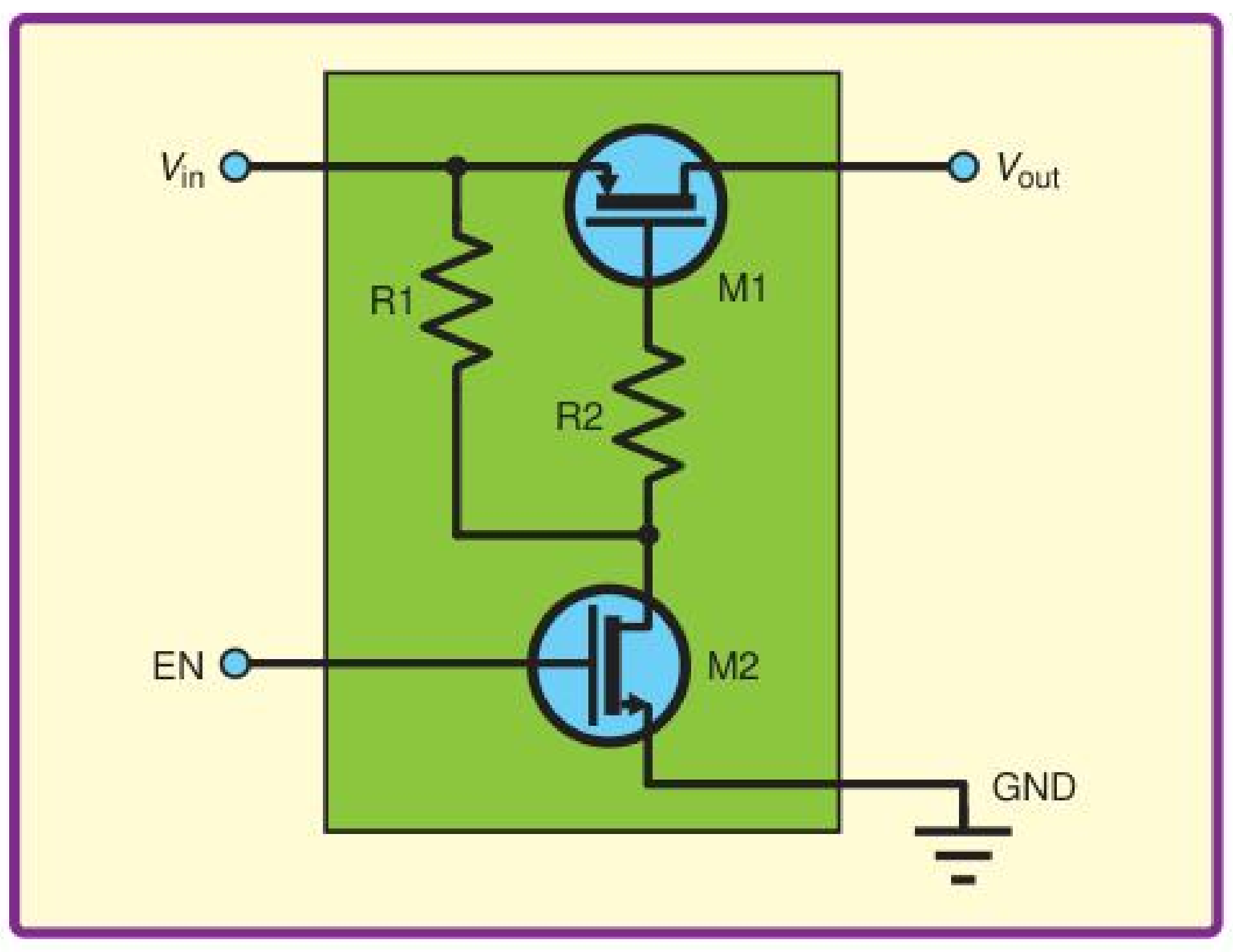

Fig.7. Load switch from Fig.1 with slew-rate-limiting resistor R2 added.

A MOSFET gate is a capacitance, and the resistor simply increases the time it takes to charge up during switch on. The effect is to relatively gradually reduce the on resistance, which causes the load capacitance to charge more slowly, reducing the inrush current. The resulting waveforms are shown in Fig.8.

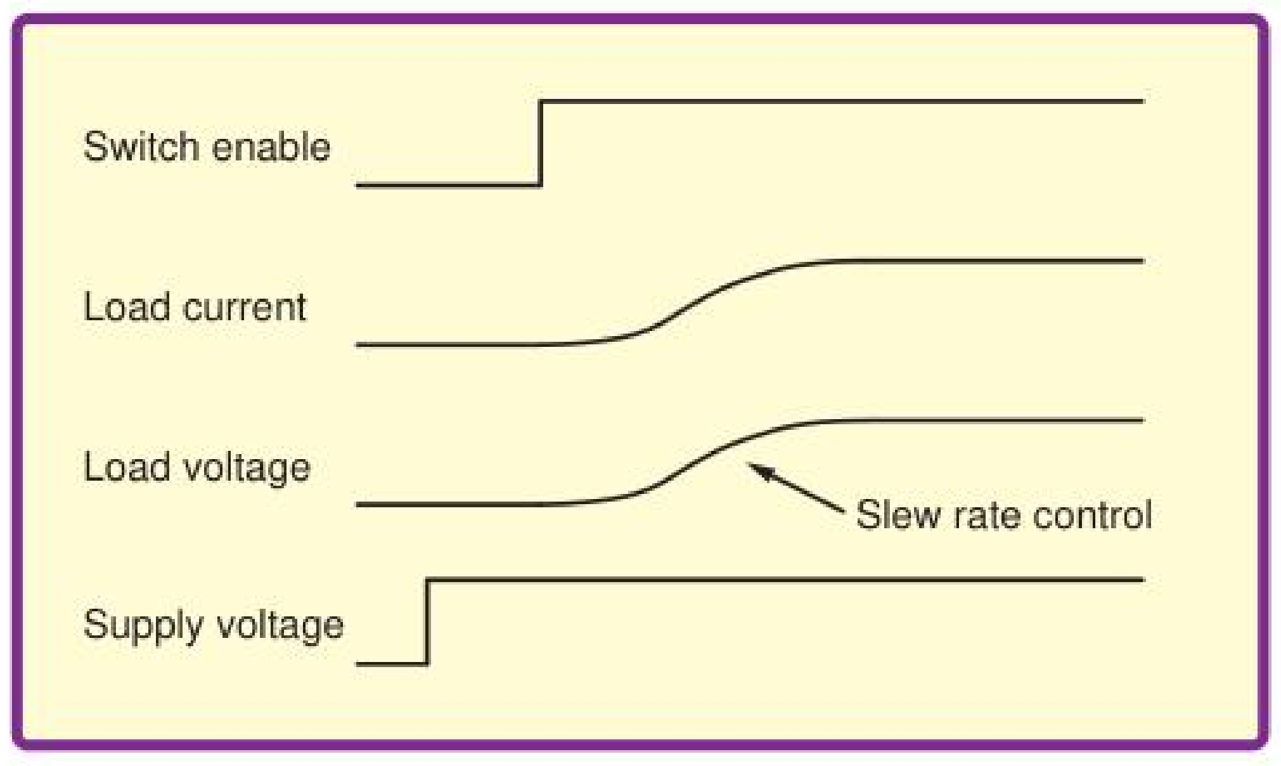

Fig.8. Load switch waveforms with slew-rate limiting (compare with Fig.6).

Rapid output discharge

There may also be problems when a load is switched off. A load switch such as the circuit in Fig.1 simply disconnects the supply. However, the disconnected circuit may continue to be active for a while, powered from the capacitance across its supplies. Again, this may occur in a simple all-on/all-off system, but will not usually have any consequences. However, if a subsystem is powered down in this uncontrolled way while other parts of the system are active, its continued, possibly erratic operation may cause problems. The solution is to switch a resistance across the load switch output to rapidly discharge a load capacitance and ensure a quick and clean shutdown.

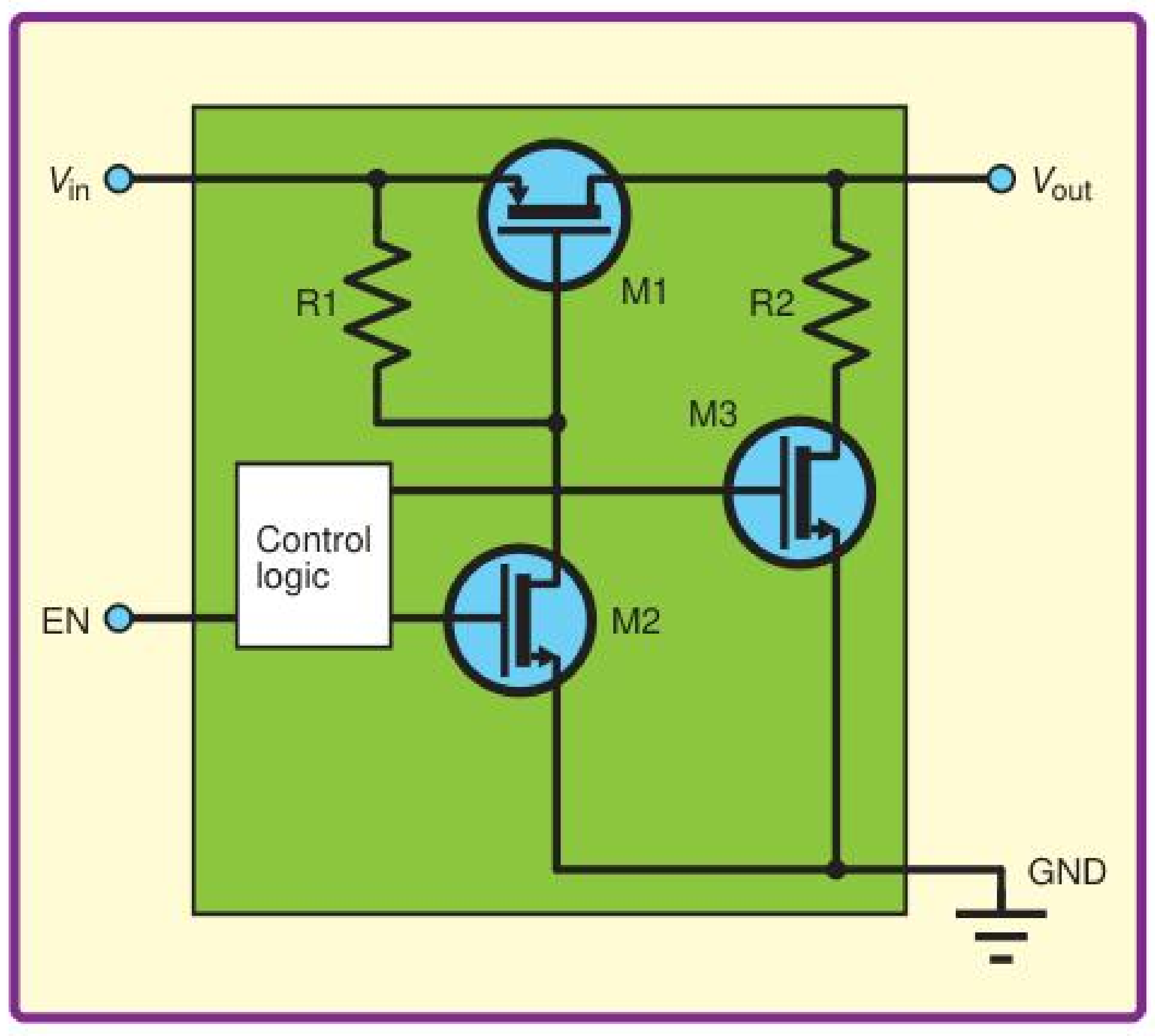

Fig.9. Load switch with output discharge.

An example of a rapid output discharge switch, applied to the circuit in Fig.1, is shown in Fig.9. The control logic switches M3 on when the load is switched off. The value of R2 is selected to provide rapid reduction in the load supply voltage without causing excessive current flow. Use of a rapid output discharge switch is not possible in all situations, for example where load switches are used to select between different input supplies (see Fig.10), or when a battery is connected across V_{OUT}. In such cases, the rapid discharge circuit would apply an excessive load to the other power source.

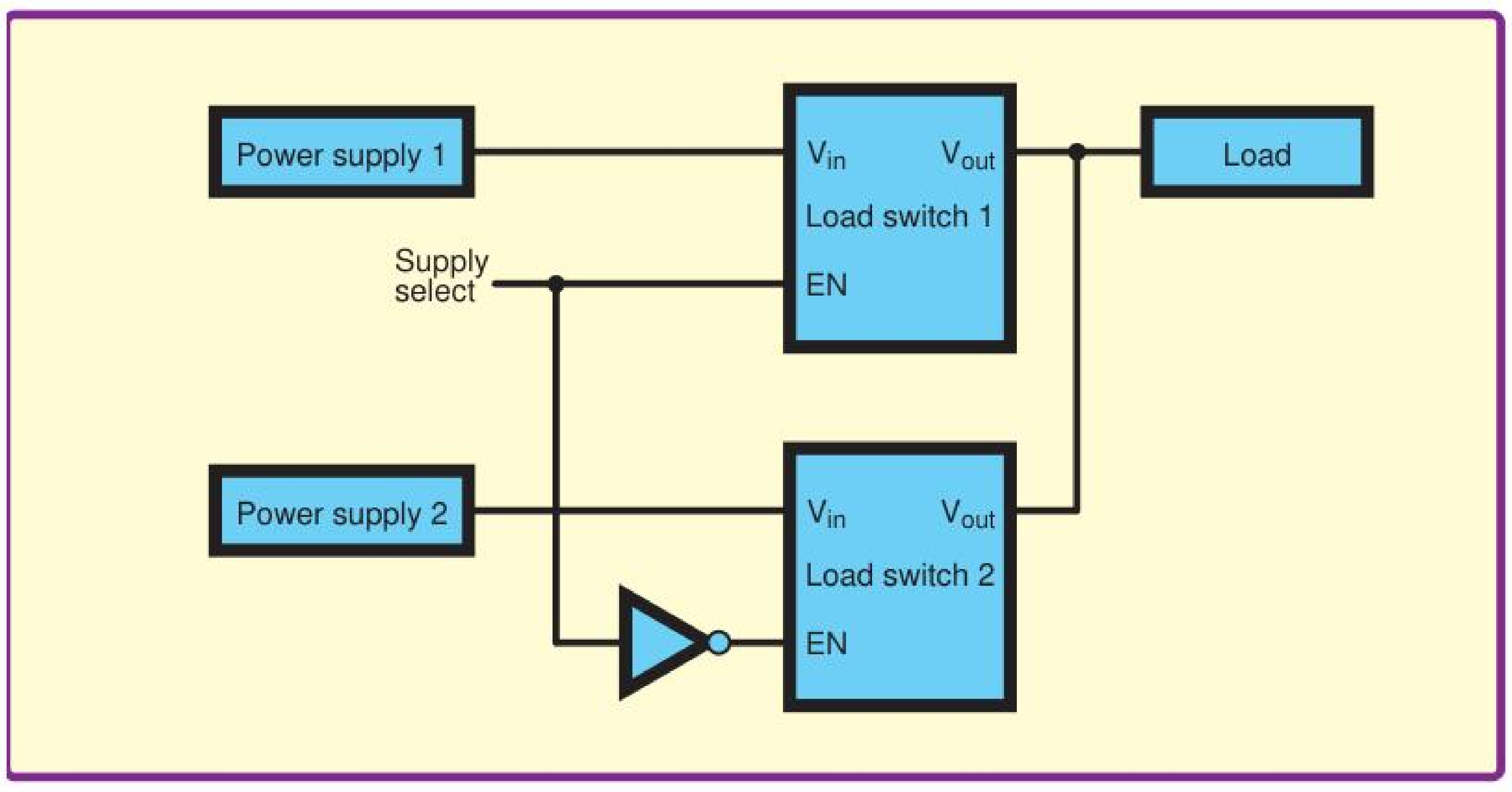

Fig.10. Supply multiplexing using load switches.

Reverse protection

In situations where a load switch is off, but a voltage is present on V_{OUT} – for example, in the supply multiplexer shown in Fig.10 – there is a possibility that a reverse current may flow through the off switch if its input voltage is less than V_{OUT}. The reverse current can flow through the body diode of the switch MOSFET and may cause damage to the transistor and other parts of the system. There are a number of ways in which load switches can be protected from reverse currents. A diode in the path from V_{IN} to V_{OUT} is a possibility, but this will drop voltage and dissipate power, which is not ideal.

Another approach to reverse-current protection is to use two MOSFETS in the switch, with their sources and drains in opposite directions, as shown in Fig.11 (for an NMOS switch). When the MOSFET is on it is effectively symmetrical in this situation – the low on resistance means that the voltage drop from source to drain is small, so the gate-source and gate-drain voltages are more or less equal. When the device is off the body diode can conduct in one direction; using two MOSFETS, as in Fig.11, means that the two body diodes in series are in opposite directions and cannot provide a conducting path. With all else equal, this circuit will have a higher on resistance than one using a single transistor, but is likely to have lower impact than using a diode in the power line.

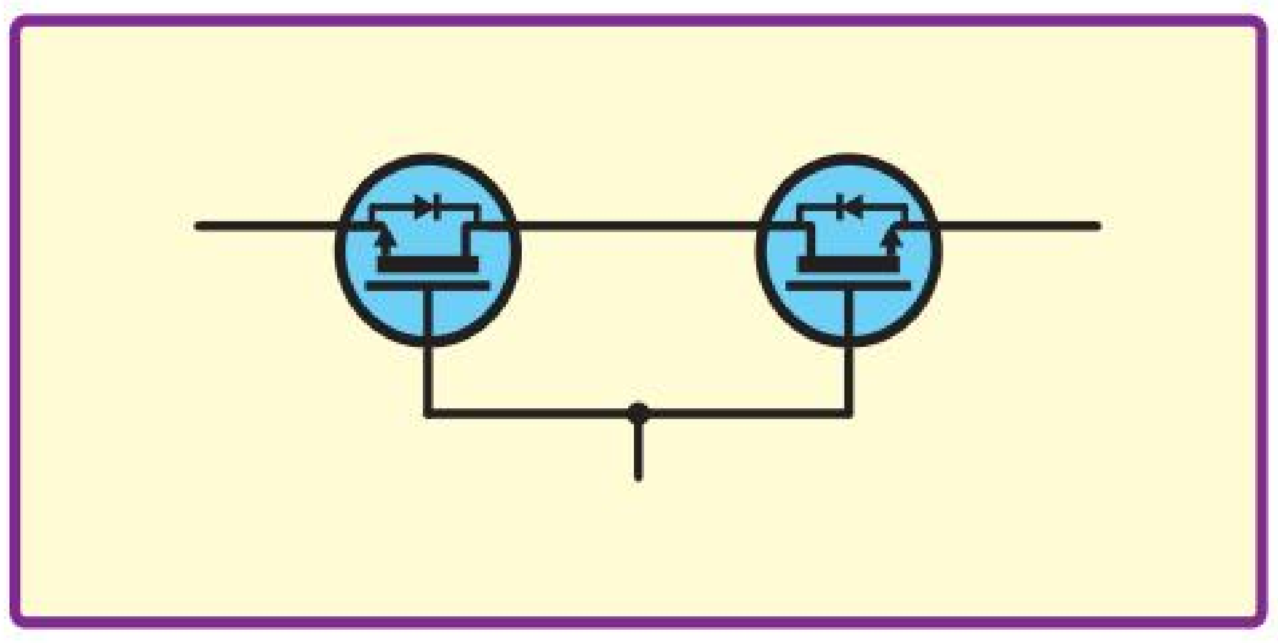

Fig.11. Two MOSFETs forming a switch with reverse-current protection.

Load switch ICs

As mentioned earlier, there are a large number of load switch ICs available. Fig.12 shows the block diagram for a pair of devices, the TPS22954/TPS22953 14mΩ on-resistance load switch from Texas Instruments, chosen somewhat randomly to illustrate some of the features that these chips offer. The two chips provide the option of either quick discharge or reverse block, as just discussed. These chips can handle V_{IN} from 0.7V to 5.7V, and currents up to 5A.

The TPS22954/TPS22953 provides output slew-rate control via a capacitor connected to the CT pin. The device

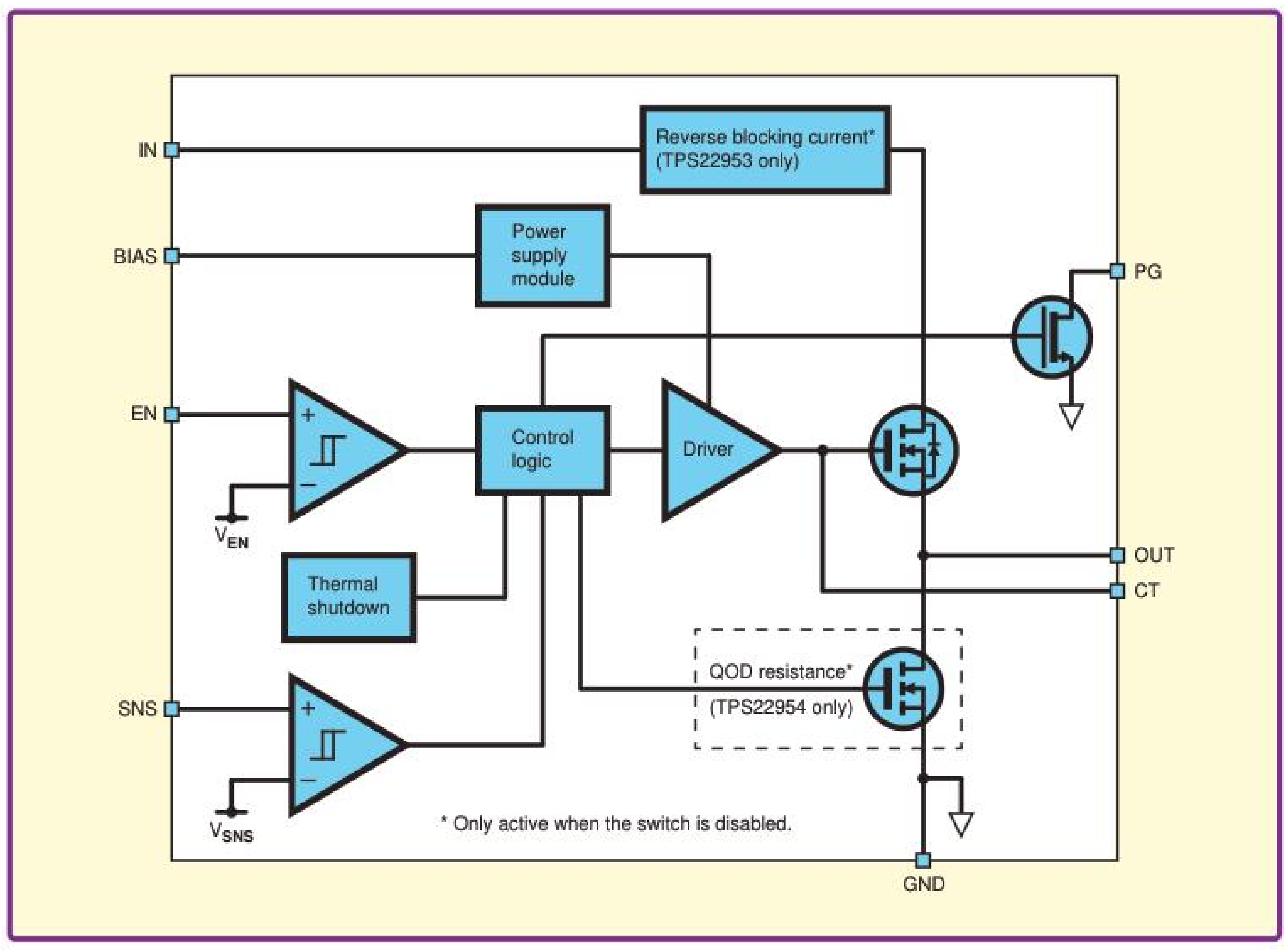
IN
Reverse blocking current* (TPS22953 only)
BIAS
Power supply module
PG
EN
Control logic
Driver
V_{EN}
OUT
CT
Thermal shutdown
QOD resistance* (TPS22954 only)
SNS
V_{SNS}
* Only active when the switch is disabled.
GND

Fig.12. Example load switch IC – the Texas Instrument TPS22954/TPS22953 (block diagram from device datasheet).

requires a separate power supply via the BIAS pin, which may be connected to V_{IN} if the voltage levels are suitable. The control logic is a little more complex than the examples discussed above. This sense pin (SNS) can be used to monitor a voltage level and provide a 'power good' output via the PG pin. One possible use for this is in power supply sequencing – once the first supply is fully on this is sensed by its load switch and its PG output is used to enable the next supply in the sequence. This is far from the full story of applications for, and features of, this and other similar ICs – consult their datasheets for details.

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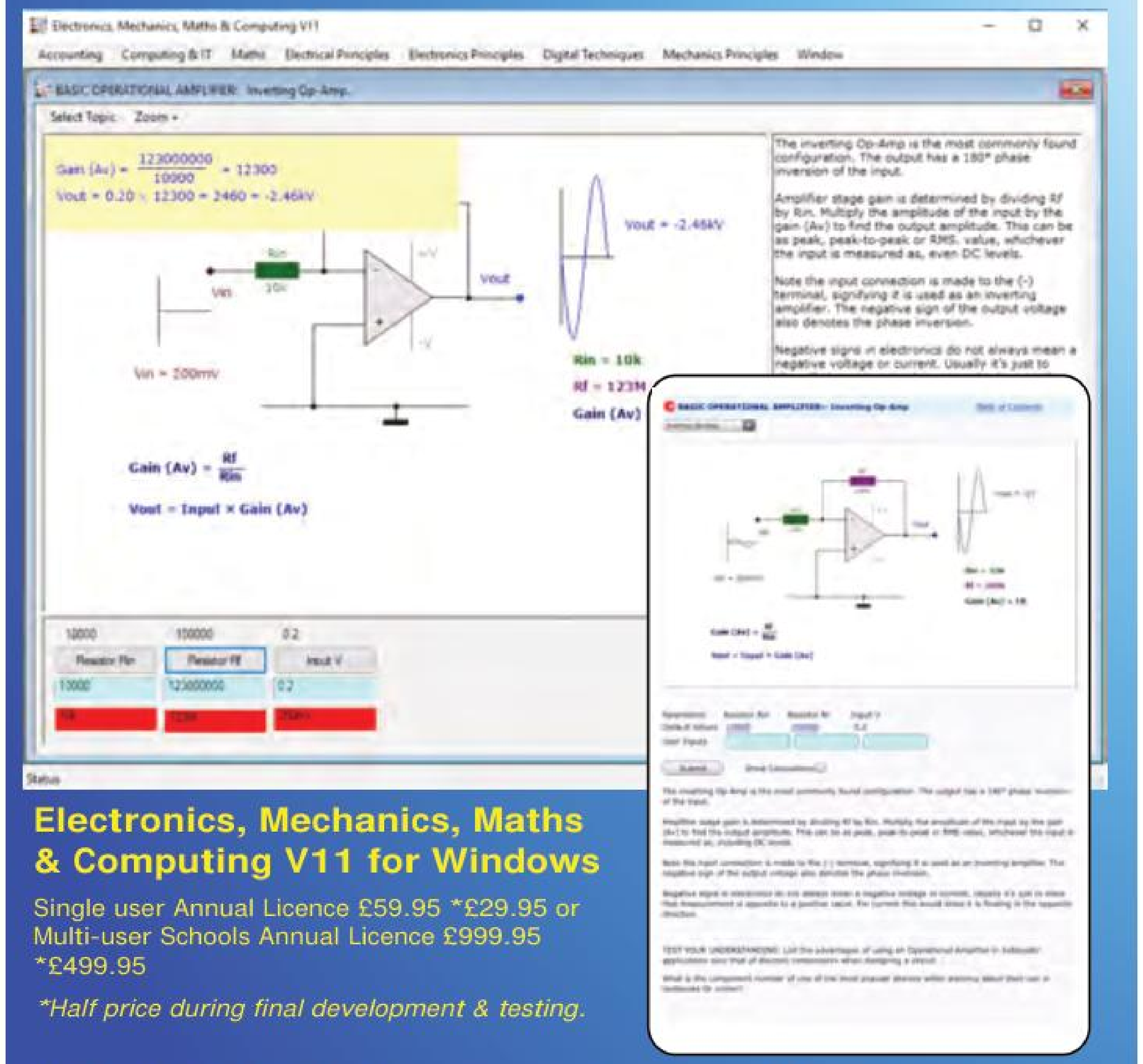

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10k
Vin
+V
-V
Vout
Vout = -2.46kV
Vin = 200mV
Rin = 10k
Rf = 123M
Gain (Av)
Resistor Rin
Resistor Rf
Input V
Status

Gain (Av) = 123000000 / 10000 = 12300
Vout = 0.20 × 12300 = 2460 = -2.46kV

Gain (Av) = Rf / Rin
Vout = Input × Gain (Av)

The inverting Op-Amp is the most commonly found configuration. The output has a 180° phase inversion of the input.

Amplifier stage gain is determined by dividing Rf by Rin. Multiply the amplitude of the input by the gain (Av) to find the output amplitude. This can be as peak, peak-to-peak or RMS. value, whichever the input is measured as, even DC levels.

Note the input connection is made to the (-) terminal, signifying it is used as an inverting amplifier. The negative sign of the output voltage also denotes the phase inversion.

Negative signs in electronics do not always mean a negative voltage or current. Usually it's just to

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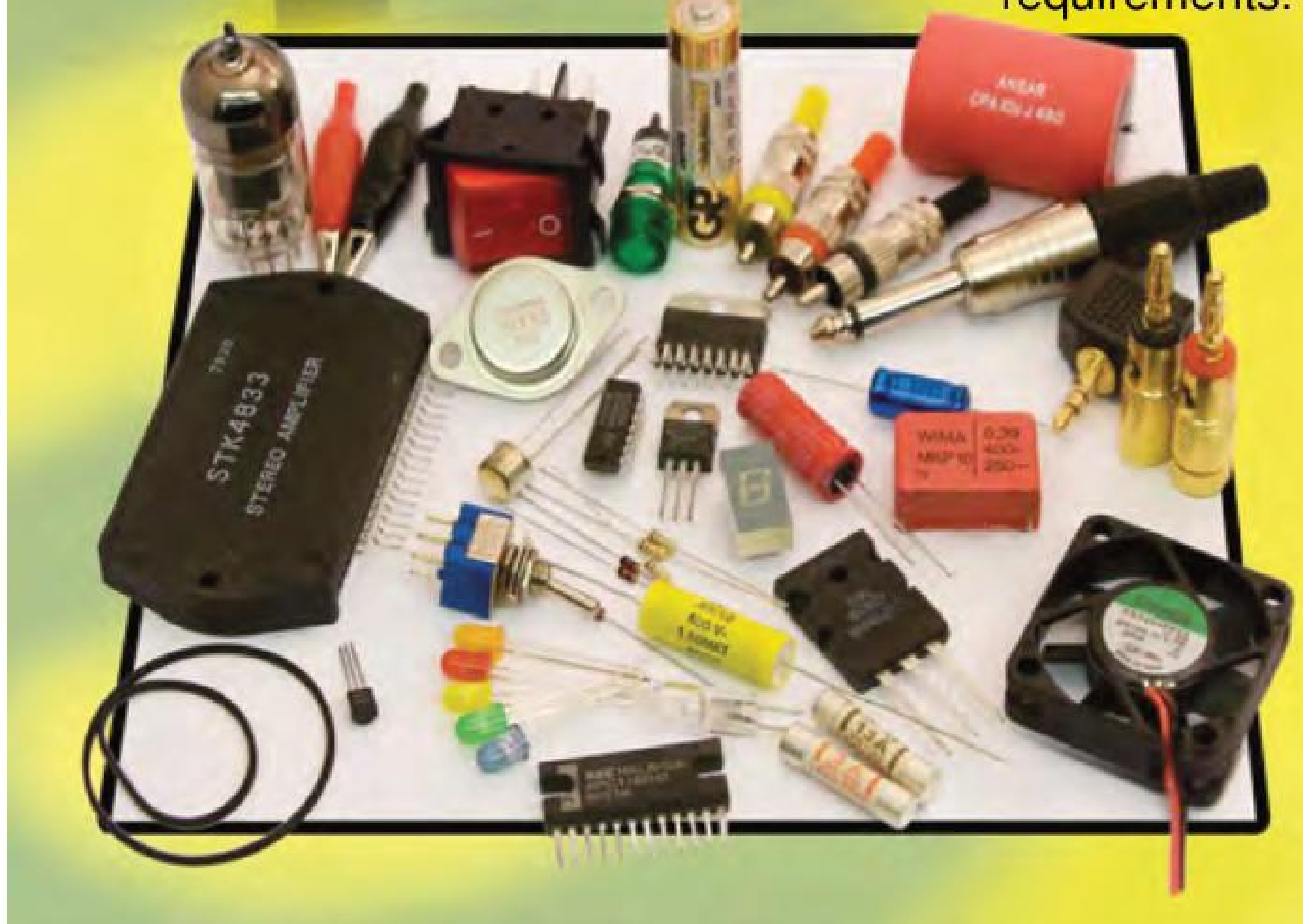

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AUDIO OUT

By Jake Rothman

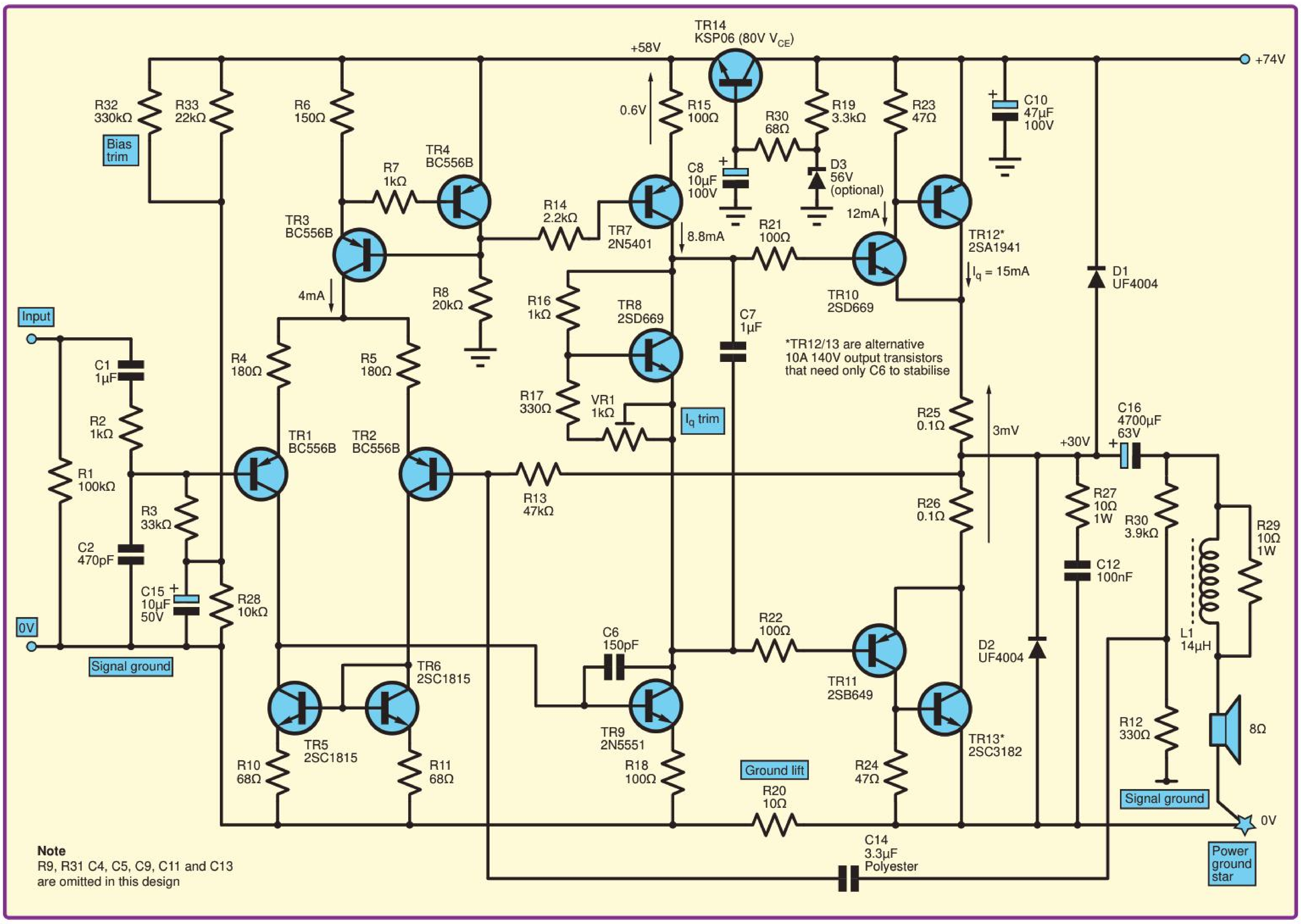
Railing against convention – Part 3

MX50 single-rail modification

It may be desirable to convert a dual-rail power amplifier design to single-rail. What better than to use the modified MX50 design described previously in *EPE*, since this is a standard circuit and the following basic modifications are universally applicable. First, an output capacitor will have to be added, the ground and negative rail will have to be joined together, and half-rail bias will have to be applied. The new single-rail power supply voltage will have to be the total of both the original rails. Instead of ±37V, the new single rail will need to be +74V.

Some internal resistors and earthing points may have to be changed. Including the output capacitor in the feedback loop should enable the original dual-rail performance over the normal audio bandwidth to be retained.

A typical value output capacitor for an 8Ω load would be 4700µF 63V, a standard low-cost electrolytic. Negative feedback is taken after the capacitor via a 3.3µF film capacitor. This ensures minimal distortion and bass loss with a –1dB point of 18Hz and a –3dB point of 10.6Hz. The power loss at 20Hz is only 4W relative to 1kHz, dropping from 53W$_{rms}$ to 49W$_{rms}$ into 8Ω. The bias voltage is set by R28 and R33 and this may have to be adjusted (using R32) with a 'scope for symmetrical clipping with different power supplies and load impedances. A capacitance multiplier using a transistor (TR14) and capacitor (C8) is necessary to avoid a big 1000µF decoupling capacitor. This is because the individual current source filter capacitor previously fitted (C5) actually added ripple with single-rail power because the ground reference had changed. A suitable solid polymer capacitor for C8 is a Kemet A759 series from Mouser. The complete circuit is shown in Fig.10.


TR14
KSP06 (80V V$_{CE}$)
+58V
+74V
R32
330kΩ
Bias trim
R33
22kΩ
R6
150Ω
0.6V
R15
100Ω
R30
68Ω
R19
3.3kΩ
R23
47Ω
C10
47µF
100V
TR4
BC556B
R7
1kΩ
C8
10µF
100V
D3
56V
(optional)
R14
2.2kΩ
12mA
TR3
BC556B
TR7
2N5401
8.8mA
R21
100Ω
TR12*
2SA1941
I$_q$ = 15mA
D1
UF4004
4mA
R8
20kΩ
R16
1kΩ
TR8
2SD669
TR10
2SD669
C7
1µF
Input
C1
1µF
R4
180Ω
R5
180Ω
R17
330Ω
VR1
1kΩ
I$_q$ trim
R25
0.1Ω
3mV
+30V
C16
4700µF
63V
R2
1kΩ
TR1
BC556B
TR2
BC556B
R1
100kΩ
R13
47kΩ
R27
10Ω
1W
R30
3.9kΩ
R26
0.1Ω
R29
10Ω
1W
R3
33kΩ
C2
470pF
C12
100nF
C15
10µF
50V
R28
10kΩ
0V
Signal ground
R22
100Ω
C6
150pF
L1
14µH
D2
UF4004
TR6
2SC1815
TR11
2SB649
TR5
2SC1815
TR9
2N5551
R12
330Ω
8Ω
TR13*
2SC3182
R10
68Ω
R11
68Ω
R18
100Ω
Ground lift
R24
47Ω
R20
10Ω
Signal ground
0V
C14
3.3µF
Polyester
Power ground star

*TR12/13 are alternative 10A 140V output transistors that need only C6 to stabilise

Note
R9, R31 C4, C5, C9, C11 and C13 are omitted in this design

Fig.10. The MX50 amplifier converted to single-rail with negative feedback around the output capacitor. Output was 53W$_{RMS}$ into 8Ω at 1kHz. At 20Hz maximum output is only reduced to 49W. An important warning is merited here: if the output of any single-rail amplifier is shorted to ground before the output capacitor, then the upper output transistor may be destroyed – be careful with those scope probes!

Compared to the dual-rail MX50, the sound is identical until clipping occurs on electronic percussive bass sounds, where a characteristic 'capacitor-induced intermodulation mushy effect' is evident. Acoustic and classical music is unaffected. Recovery from clipping is faster on DC-coupled amplifiers. Also, the beneficial effect of feedback around the output capacitor is lost when the feedback loop is broken at clipping.

Regulation of the low-power stages supply and bias can be done by inserting a 56V Zener diode into the capacitance multiplier across C8. Using simple potential divider bias, rock-solid low-frequency stability is obtained, although output power is reduced from 53W to 46W into 8Ω (at 1kHz). I made the Zener diode switchable for comparison. On listening, I found the reduction in output power outweighed the improvement in stability. The amp remained cleaner to a louder volume without the Zener diode. What looks better on the 'scope does not necessarily sound better. Fig.11 shows the prototype single-rail MX50.

Other amplifiers

Applying output capacitor negative feedback to older amplifiers with a single-transistor input stage (as opposed to a long-tailed pair) is more complicated. A significant current is normally required to pass through the feedback resistor to power the input transistor. This necessitates a low value, typically a few kΩ, which means little capacitor feedback can be applied. The DC feedback resistor can be increased up to 1MΩ by employing an extra transistor as a current booster, thus enabling more AC feedback, as shown in Fig.12. The higher impedance also means the film coupling capacitor value can be minimised. Care has to be taken to reduce the low-frequency open-loop gain elsewhere to avoid low-frequency response humps and consequent 'bouncing' on transients. In this case, there was a peak of 3dB at 8Hz, cancelled out with a smaller input capacitor.

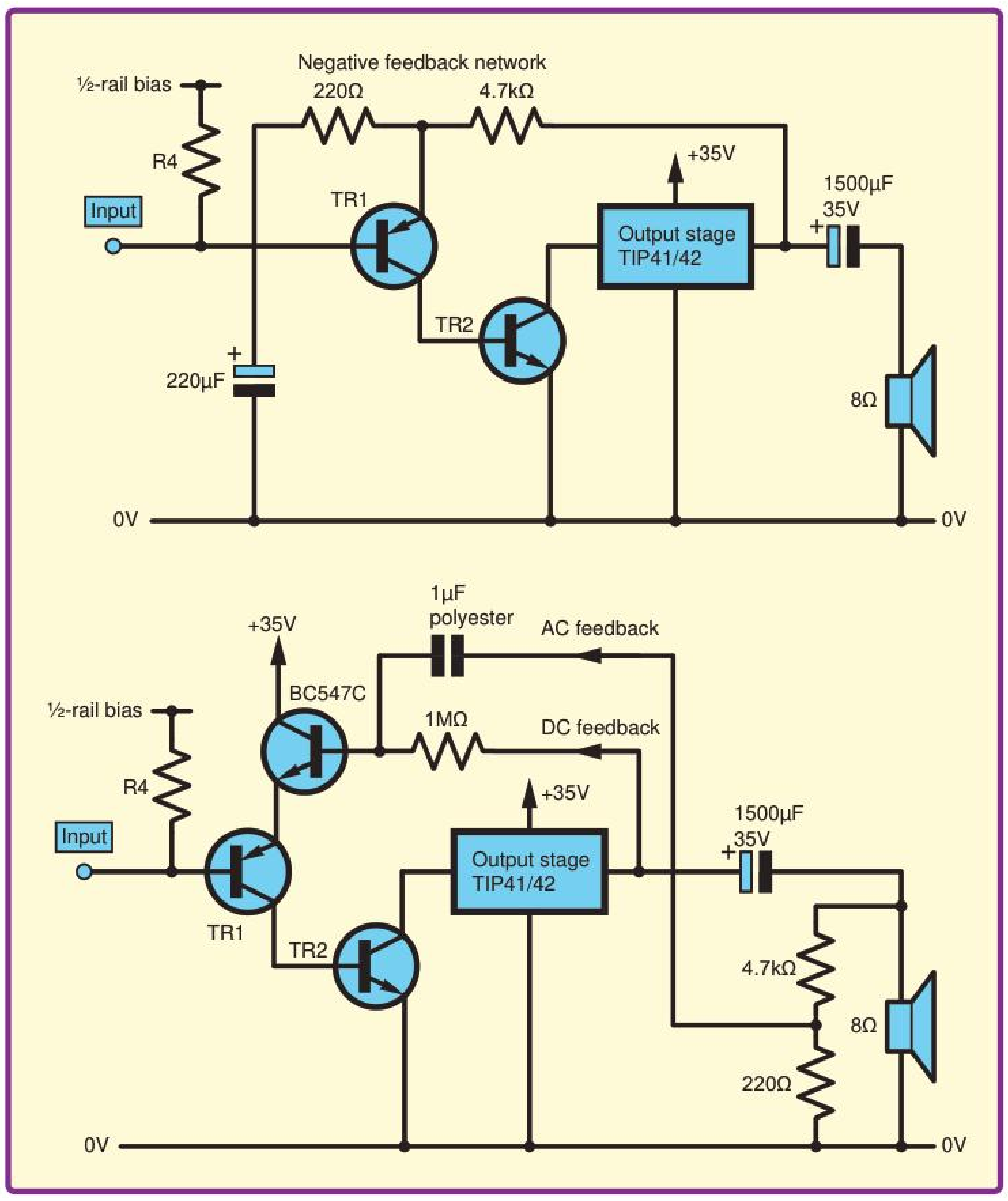


*Fig.12. Applying output capacitor negative feedback to a single input transistor amplifier. In this case the Super Simple Retro design (*EPE *Sept 2016, Fig.23).*

MX50 output capacitor voltage rating

Even though the power supply of the single-rail MX50 is 74V it is possible to get away with 63 or even 50V. This penny-pinching can't be taken too far however. I have seen amplifiers where the output capacitor voltage rating is half the supply voltage, which is dangerous. If the top transistor in a push-pull output stage short circuits and the bottom transistor fails open, then the full rail voltage will be applied to the capacitor. The fuse will blow eventually, but the capacitor may short-circuit, explode and cause speaker burn-out. In practice, however, it does seem to be safe to rate it a bit lower. Hundreds of Leak Delta 70 amplifiers used output capacitors rated at 50V with a 75V rail with no failures in the field. Also, Roberts radios used 6.3V capacitors with 9V supplies. I expect this was because the amplifier's output never quite reached full rail voltage. Possibly, if it did, it only remained high for brief periods, insufficient for the capacitor to fully charge up. Pulses would also have been easily accommodated by the capacitors surge rating, which is generally 20% above the rated voltage.

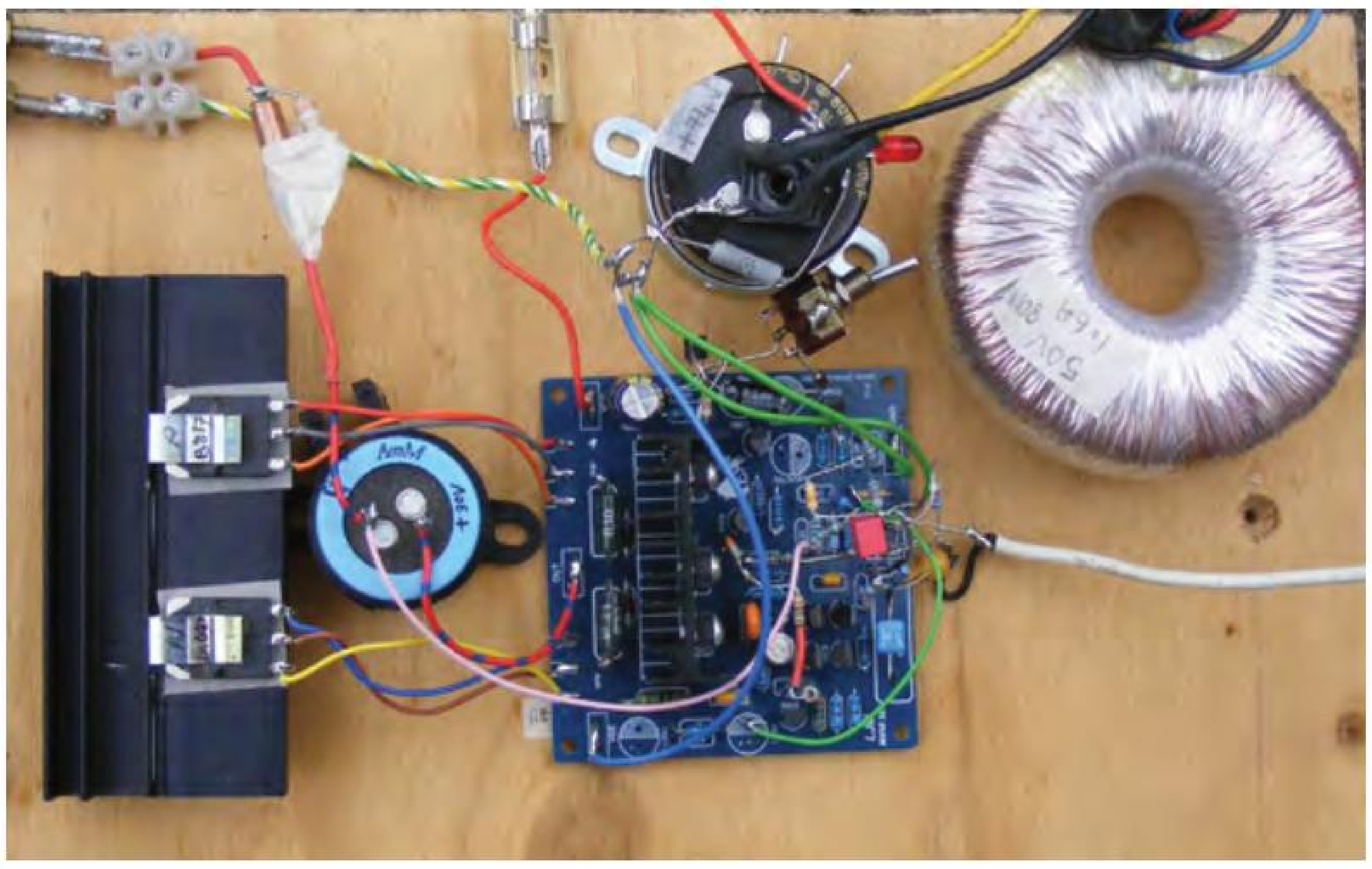


Fig.11. A prototype of the single-rail MX50. Note the blue output capacitor and the single smoothing capacitor. Only one fuse is needed, a bonus of single-rail design since fuses are unreliable.

For the single-rail MX50 I used a 50V (60V surge) output capacitor with an 80V 6800µF smoothing capacitor. In the single-rail MX50 circuit I ended up with only three wet electrolytic capacitors for a stereo amplifier, compared to my absolute minimum of two for the dual-rail version (*EPE*, Dec 2017).

Note that two 10,000µF 50V capacitors on a dual-rail power supply store the same energy as a single 10,000µF capacitor of 100V rating on a single-rail. Output and smoothing capacitors above the standard 63V rating are required with transformers over 44V (40W into 8Ω) which cost disproportionately more.

Funny noises

On single-rail systems the output capacitor has to charge up to half-rail at turn on. If this is done too quickly, a nasty speaker thump results. This can be mitigated by slowly raising the bias or rail voltage. This also has the advantage of slowly turning the amplifier on. The value of the bias decoupling capacitor C15 in the MX50 has been specially chosen to ramp up smoothly.

A pull-down resistor of around 2.2kΩ is needed after the output capacitor to prevent a loud pop when the speaker is connected.

Summary

A single-rail design with capacitor coupling is best for small power amps. Above 40W, where clipping is likely and for those who love hard punchy electronic bass, the problems associated with output capacitors become more significant. The increased additional cost of DC protection circuitry with relays becomes proportionately less at higher powers.

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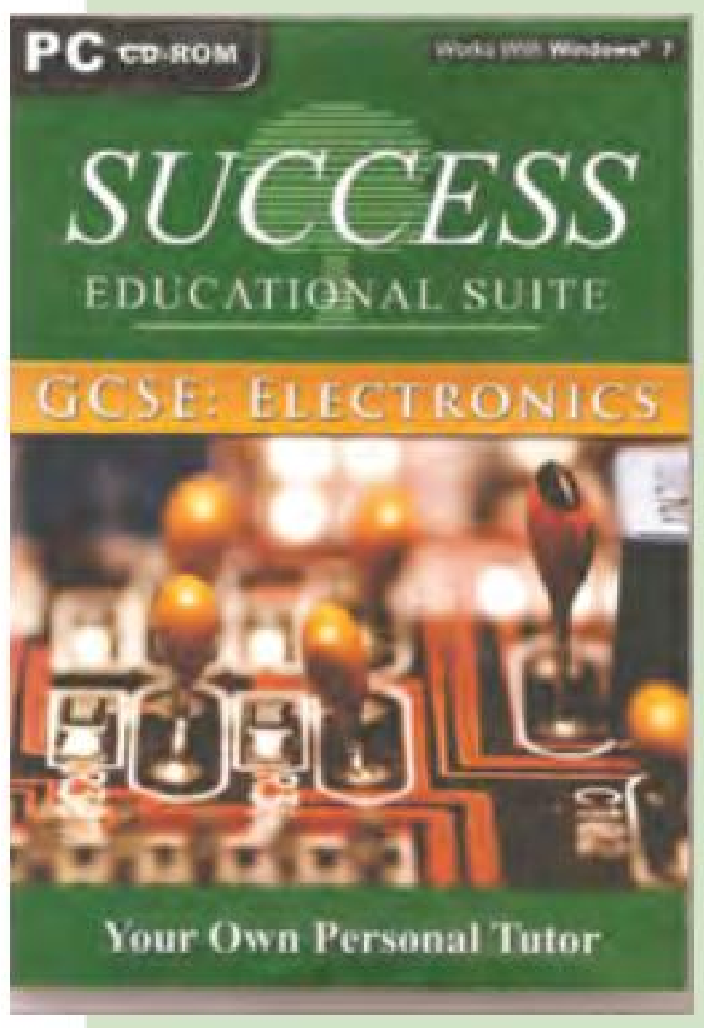

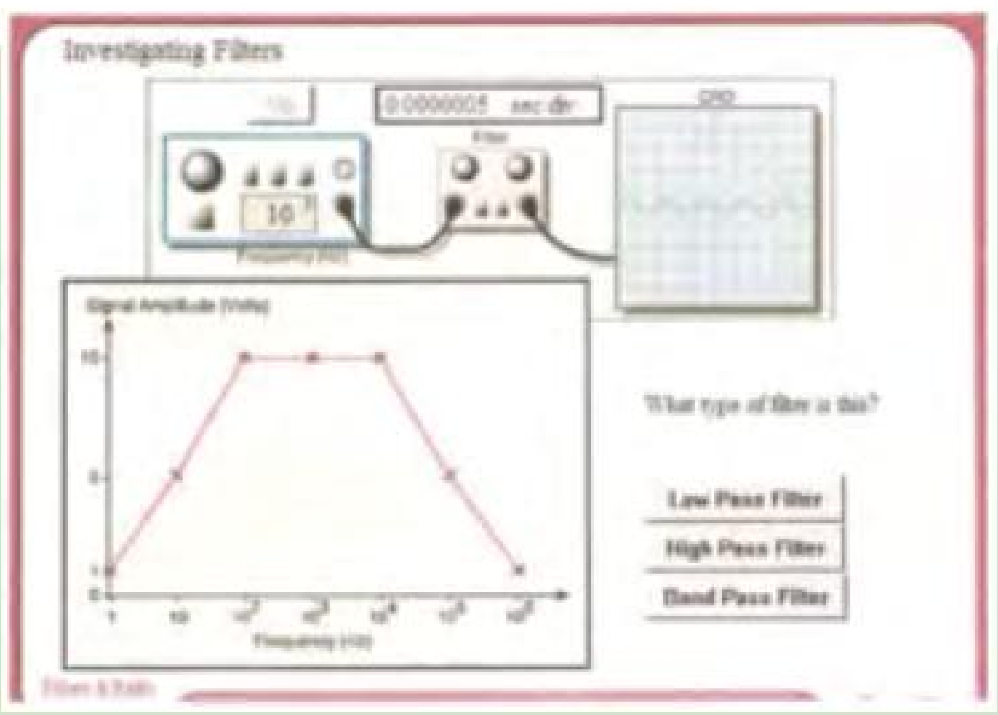

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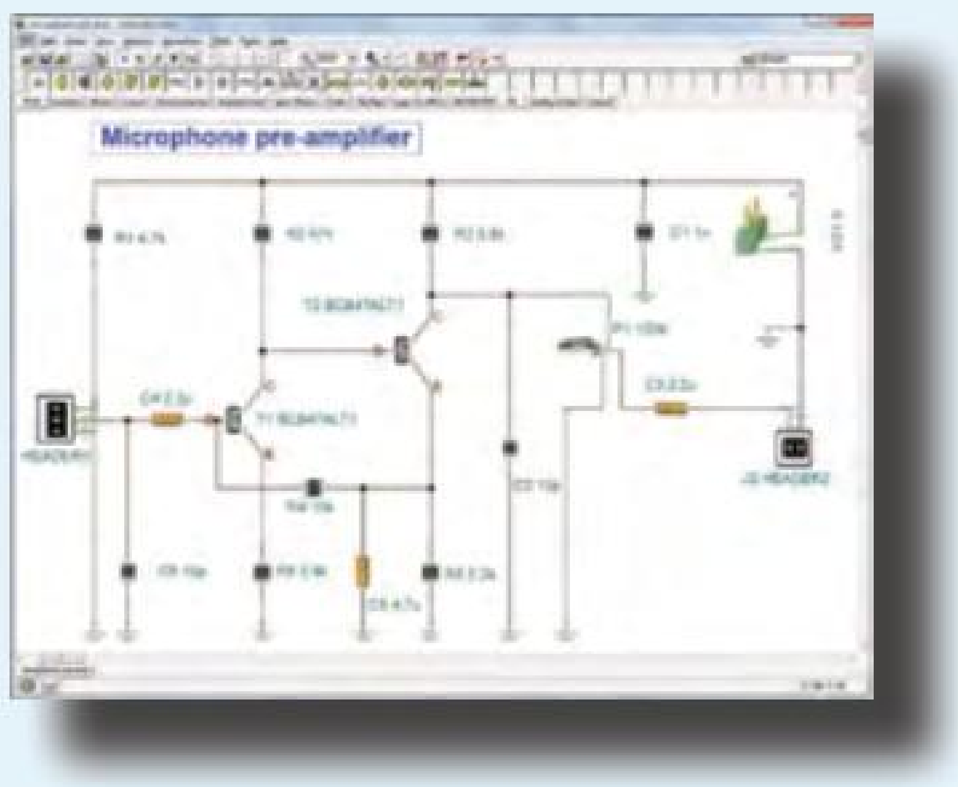

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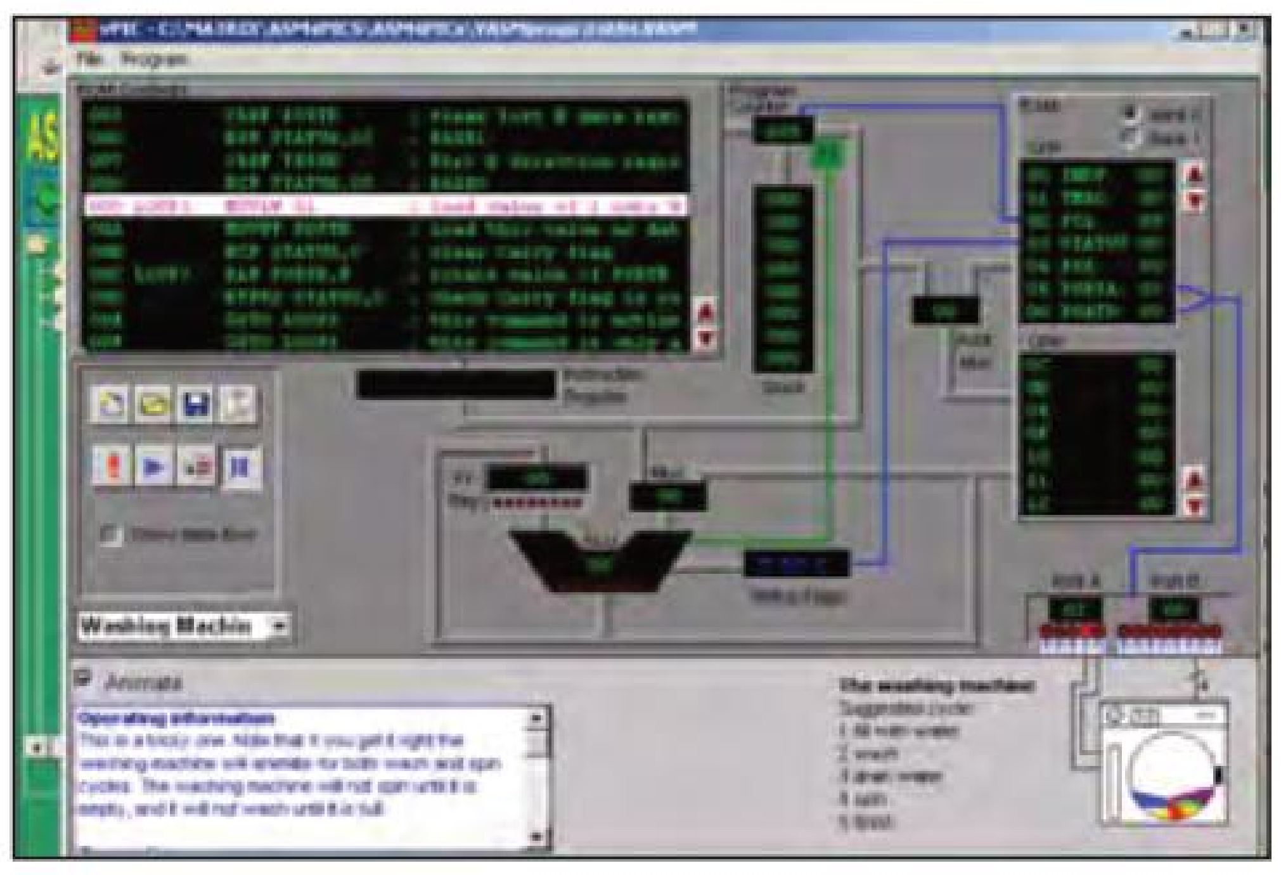

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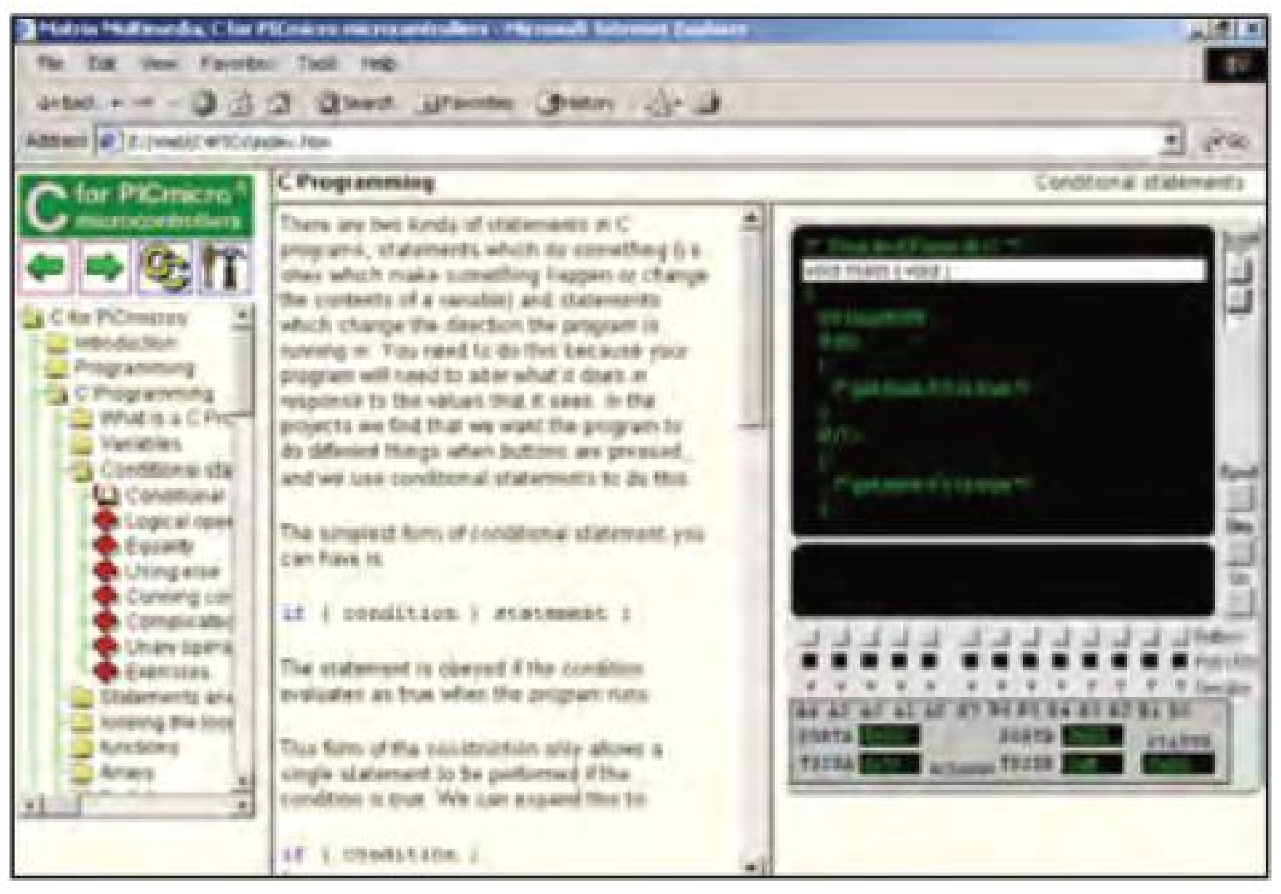

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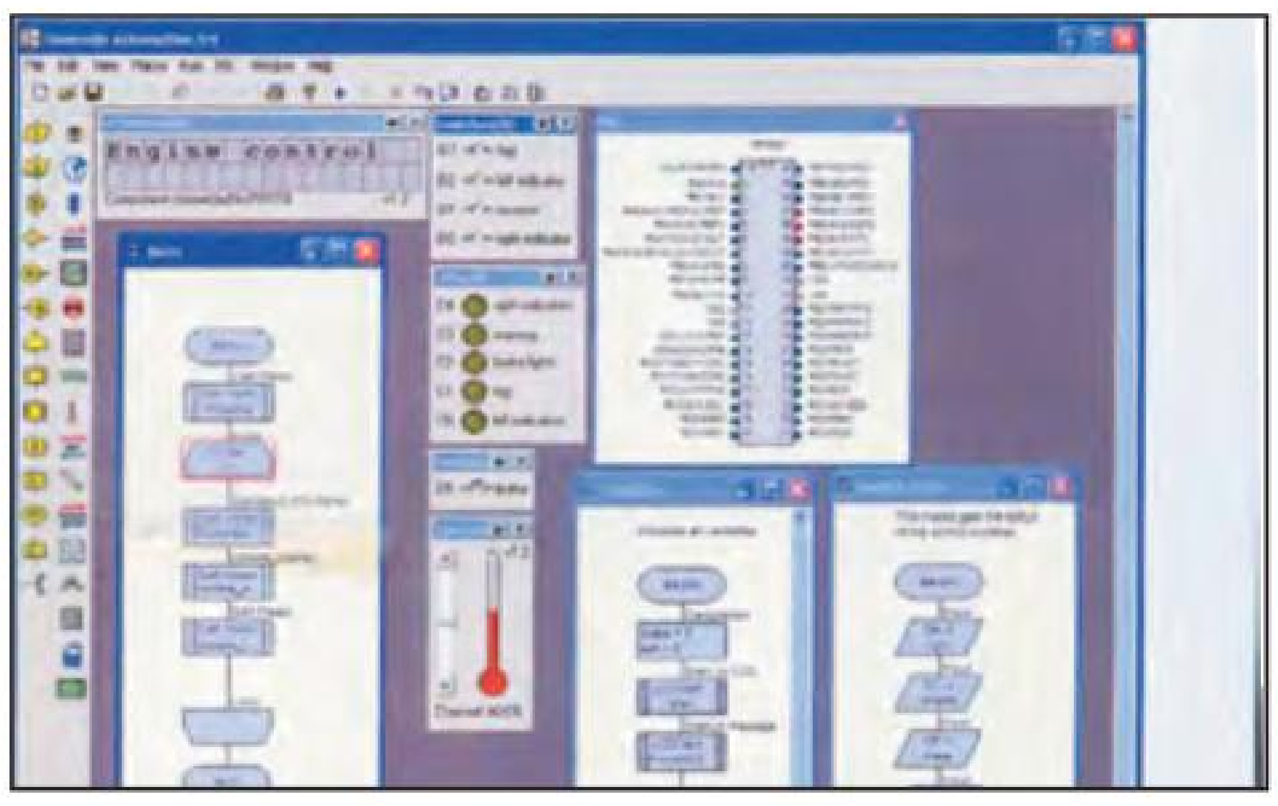

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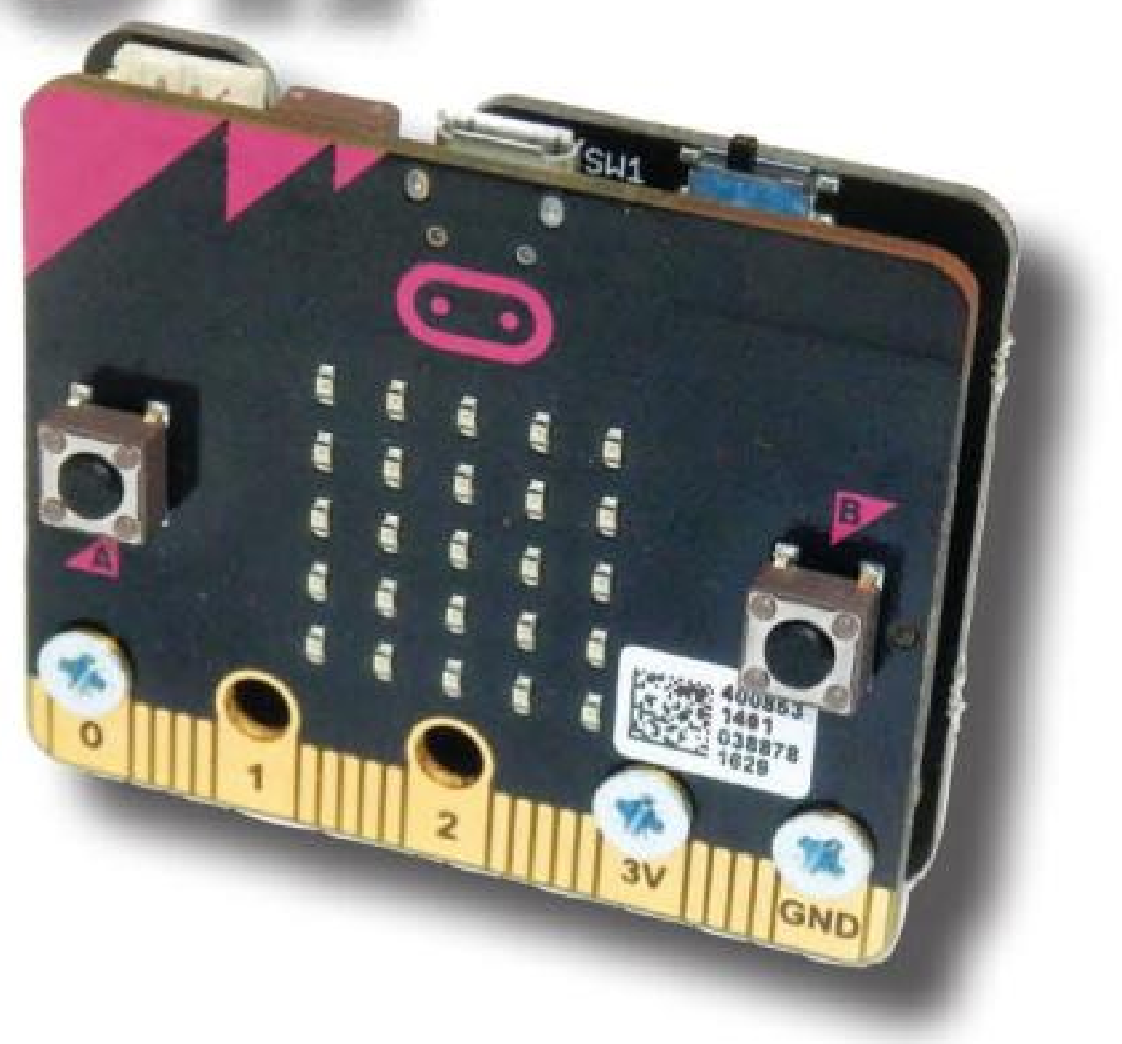

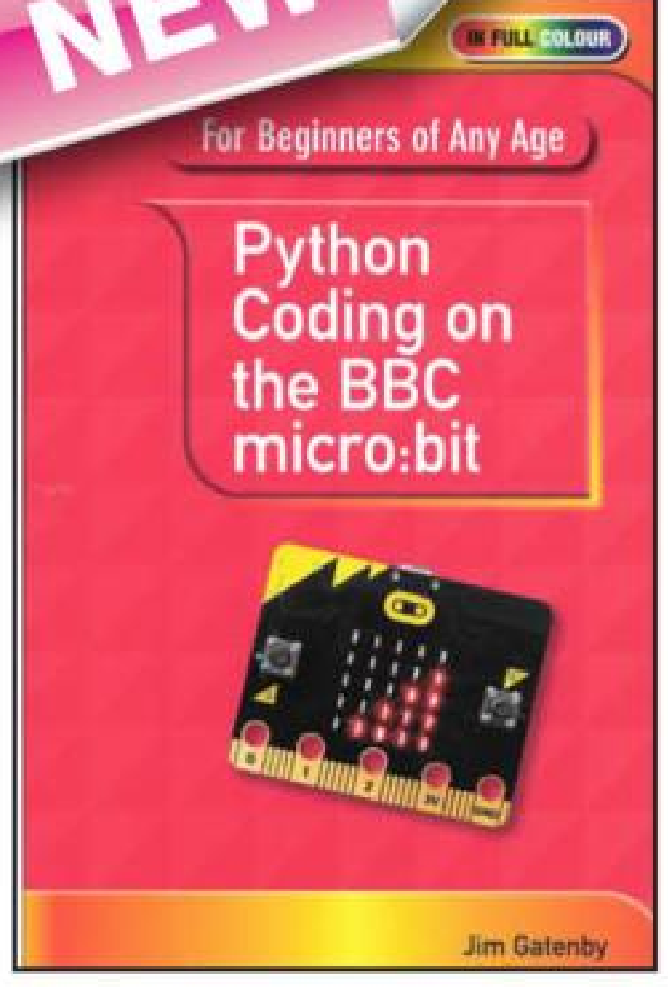

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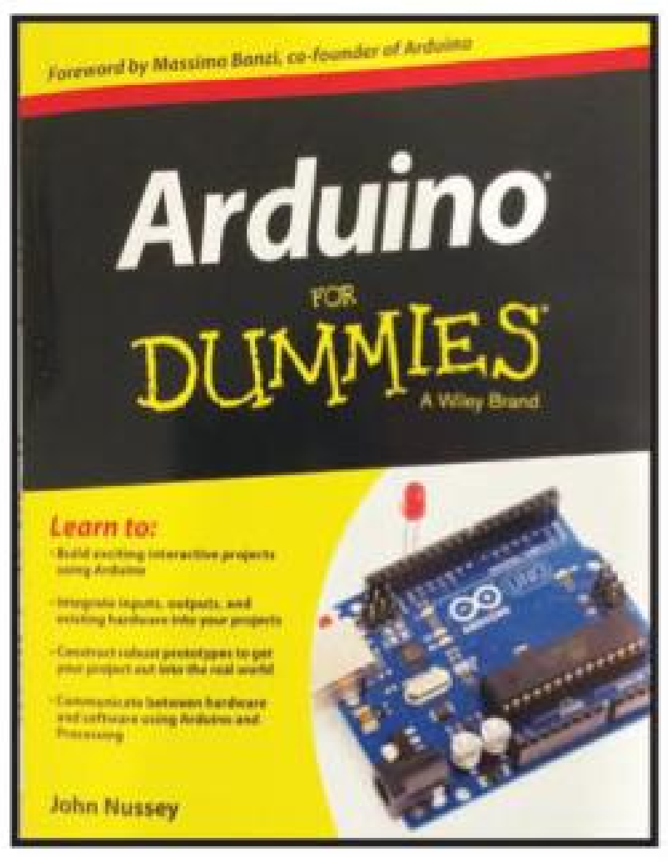
Teach-In 2016
Exploring the Arduino



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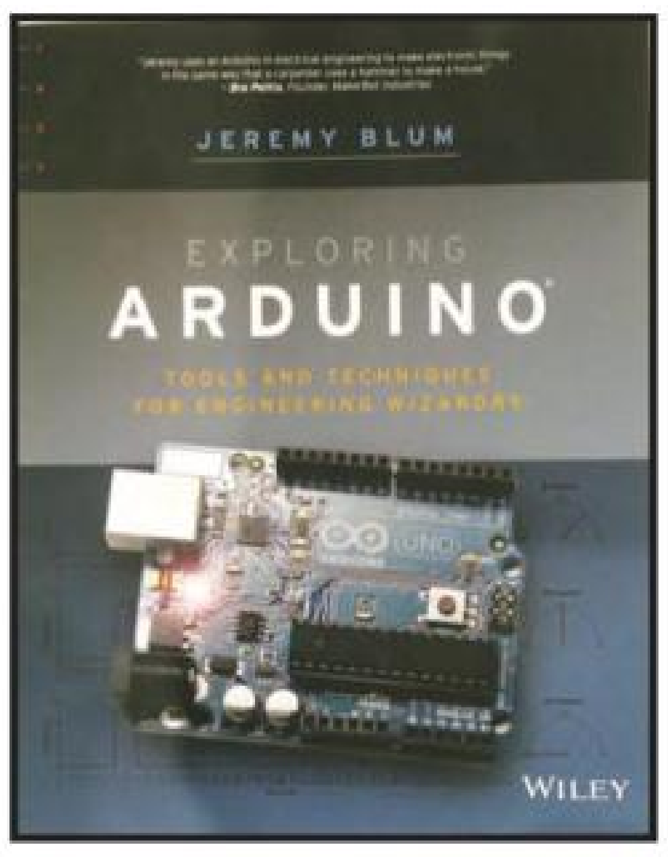

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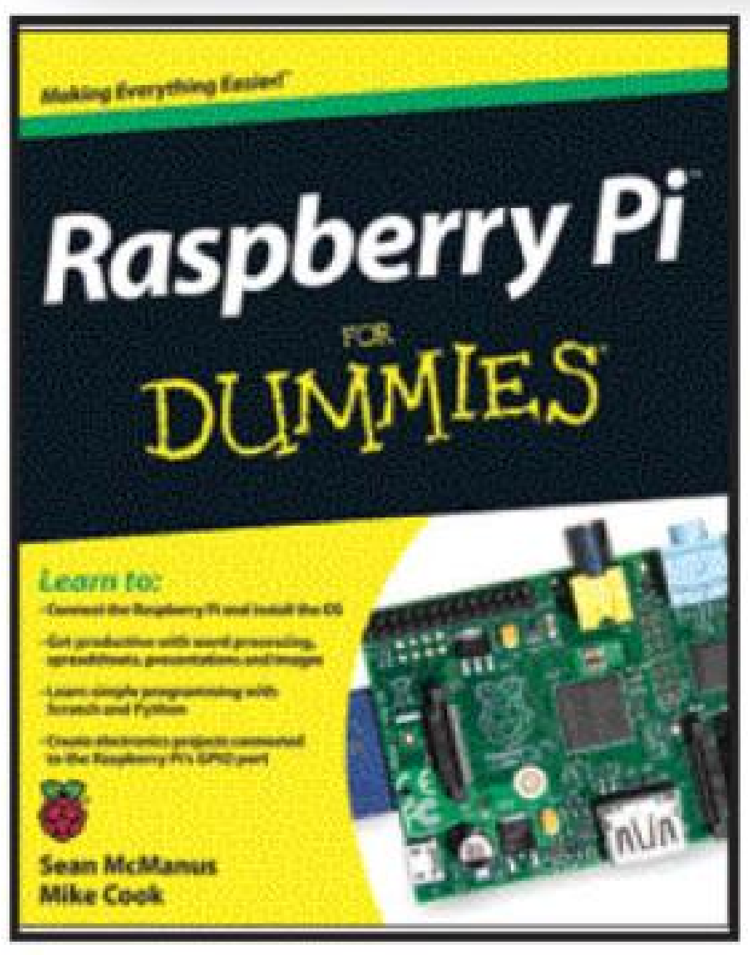

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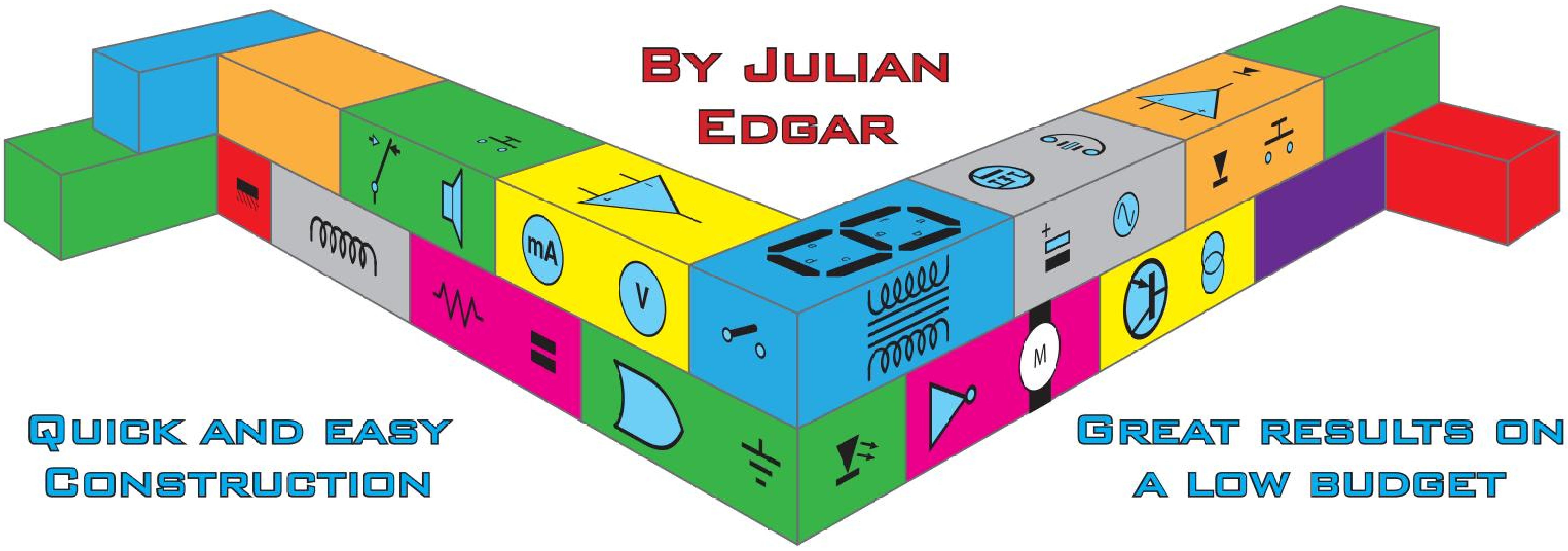

By Julian Edgar

Quick and easy Construction

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High-current flasher

Large complex projects are fun, but they take time and can be expensive. Sometimes you just want a quick result at low cost. That's where this series of *Electronic Building Blocks* fits in. We use 'cheap as chips' components bought online to get you where you want to be... FAST! They represent the best value we can find in today's electronics marketplace!

Flashers are useful in lots of situations. You might like to flash a warning light or pulse a beeper when something goes wrong. In this situation, a simple circuit will do the trick – but what if you want to flash higher current loads? Then you'll need to add a suitable switching device – and probably a heat sink as well. And what if you want a flash pattern that's more complex than just on/off/on/off? In that case, you'll need to add some kind of microcontroller.

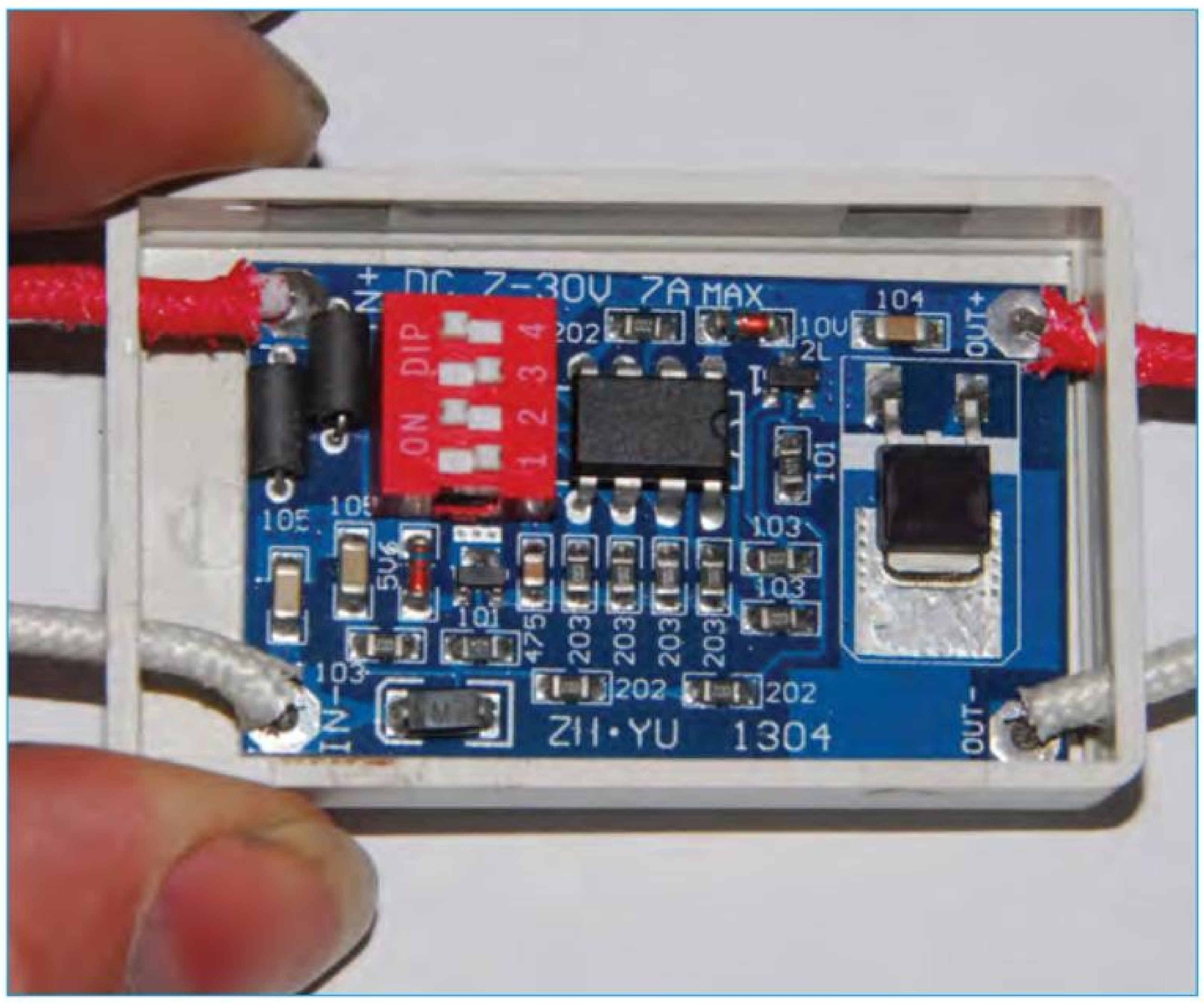
Or instead, you can just buy the multi-mode flasher module that's shown here.

At under £5 (including delivery), it's cheap enough to replace low-current applications… but with the ability to drive loads of LEDs and incandescent bulbs drawing up to 7A, and with no less than 16 different flashing modes available at the touch of a switch, it's in a league of its own in terms of bang for your buck.

While marketed as a device to flash your brake lights, the '12V-24V LED Brake Stop Light Lamp Flasher Module Flash Strobe Controller 16 Mode' module (eBay no. 282704940728) is much more than 'just' a car brake lights flasher.

Packaging and connections

The flasher comes packaged in a 58 × 35 × 16mm box – very compact indeed. At each end of the box are two wires – red and white. These are the input and output leads (red for positive). Marked on the box is an 'O' and arrow – this stands for 'output'. (If your box is not marked, the output end is that furthest from the DIP switch and closest to the output transistor.) Within the box (and accessed by lifting the lid) is a 4-position DIP switch. This switch is used to select which one of the 16 different flashing modes you want.



This tiny flasher module has 16 different modes and can handle loads of up to 7A. At under £5 delivered, it's amazing value!

Flashing modes 1-8

Let's start with the most common use first. You want to flash a high-current LED as a warning. Connect the LED (complete with appropriate dropping resistor, and observing the correct polarity) across the output leads. Set all the DIP switches to '0' (all the switches set closest to the output end) and then connect power (7 to 30V DC). The LED will flash at 18Hz (I measured 17Hz, but that's close enough). Set switch 1 to position 1, and the flash rate changes to 12.5Hz (I measured 11.8Hz). **Note that you must disconnect power and then re-apply it for the change in switch position to take effect.**

Flash rates from the aforesaid 18Hz down to 1Hz are available, all with a 50% duty cycle. These settings comprise Modes 1 to 8.

Flashing modes 9-12

The next range of modes (Mode 9 to 12) are lower-current-consumption modes. Mode 9 turns on the output for 50ms once per second (a 5% duty cycle), while Mode 10 flashes three 50ms pulses then stops for a second, before restarting the cycle. Mode 11 outputs three 50ms pulses, but this time stops for two seconds. Mode 12 flashes ten 50ms pulses then stops for two seconds.

Flashing modes 13-16

Modes 13 to 16 cause gradual changes in the LED output. The rate of brightening and darkening varies with the different modes – for example, Mode 13 varies brightness one up/down cycle every three seconds, while Mode 16 does four cycles over 11 seconds.

Table 1: High-current flasing modes (This table shows the 16 different flasher modes that can be selected by setting the on-board DIP switch.)

Mode	S1	S2	S3	S4	Output
1	0	0	0	0	18Hz flash
2	1	0	0	0	12.5Hz flash
3	0	1	0	0	9.5Hz flash
4	1	1	0	0	6.5Hz flash
5	0	0	1	0	4.5Hz flash
6	1	0	1	0	3Hz flash
7	0	1	1	0	2Hz flash
8	1	1	1	0	1 Hz flash
9	0	0	0	1	Cycling single 50ms flash, stopped 1s
10	1	0	0	1	Cycling three 50ms flashes, stopped 1s
11	0	1	0	1	Cycling three 50ms flashes, stopped 2s
12	1	1	0	1	Cycling ten 50ms flashes, stopped 2s
13	0	0	1	1	One up/down brightness cycle every 3s
14	1	0	1	1	Two up/down brightness cycles every 5s
15	0	1	1	1	Three up/down brightness cycle every 7s
16	1	1	1	1	Four up/down brightness cycle every 11s

Incidentally, this variation in brightness is achieved via pulse width modulation (PWM) at a frequency of 235Hz.

The beauty of this 'multi-mode' approach is that the flasher function can be very much tailored for the situation. For example, if current draw isn't an issue and you want to attract attention, then flash a high-power LED at 6.5Hz. But if you want minimal current draw, select a single 50ms flash every second. The 'output off' current draw of the module is only 7mA, so overall power consumption will be quite low.

Use with incandescent lamps

But what if you want to flash incandescent lamps? The rapid available flashing rates (like 9.5Hz) and short pulses (50ms) won't work with incandescent lamps – the thermal inertia of the filaments means that they just won't respond fast enough. In that situation, the 1Hz, 50% duty cycle mode (Mode 8) can be used, and the 'varying brightness' modes (Modes 13 to 16) are very effective.

And what about current handling? The module is rated at 6A on incandescent loads and 7A on LED loads. And how well does the current handling stack up? Very well, in fact. It's most likely that you'll be running high current loads when using incandescent lamps, with the most 'current hungry' mode being the cycling modes of 13 to 16. I ran the module at 8A in Mode 16 with incandescent loading, and all was fine. On LED loads, 7A is an awful lot of lighting – I ran an automotive LED light bar (7.5A) and again the module coped fine.

Compact, versatile, cheap and it does just what is says it will do – a must-have for the parts drawers!

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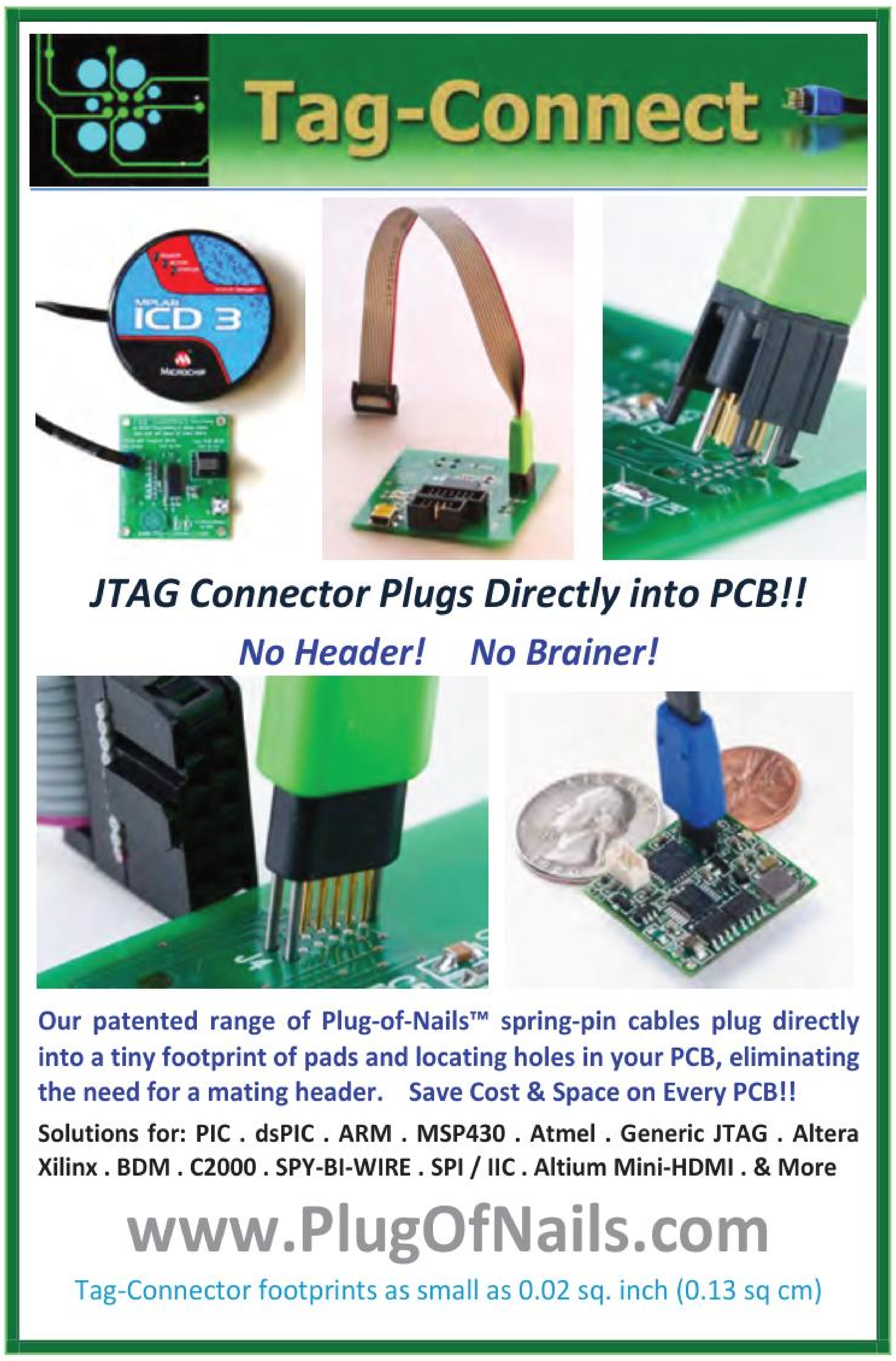

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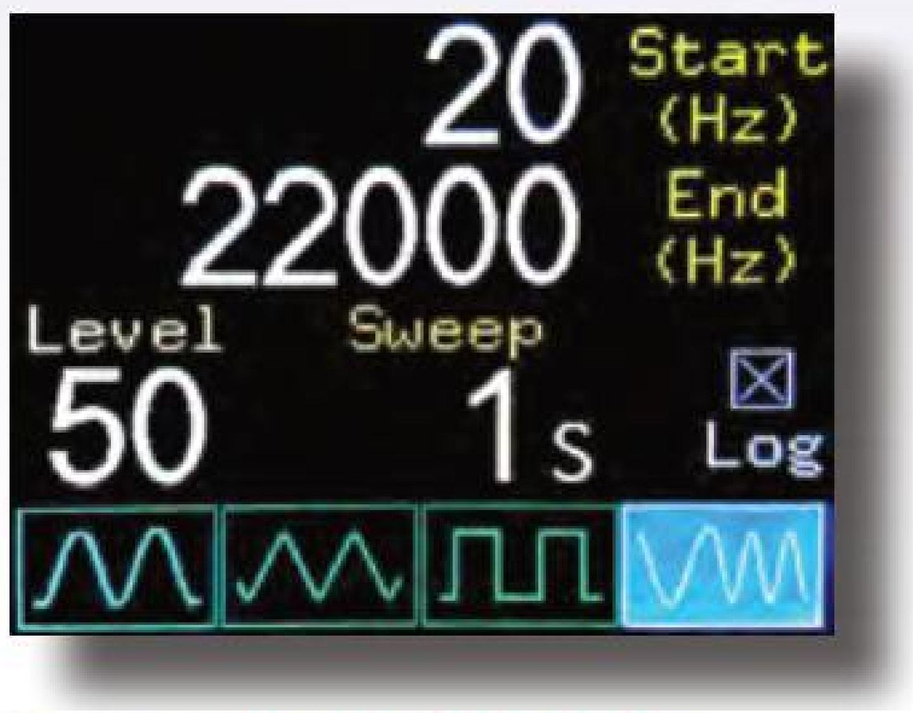
For editorial address and phone numbers see page 7

Next Month

APRIL '18 ISSUE ON SALE
1 MARCH 2018

Spring Reverberation Unit

Here's a blast from the past! Despite the availability of digital reverb and effects units these days, many musicians still like the 'old school sound' of spring reverberation. This new design uses a cheap, readily available spring 'tank' with a flexible power supply, so you can easily build it into your favourite amp, even if it's portable.

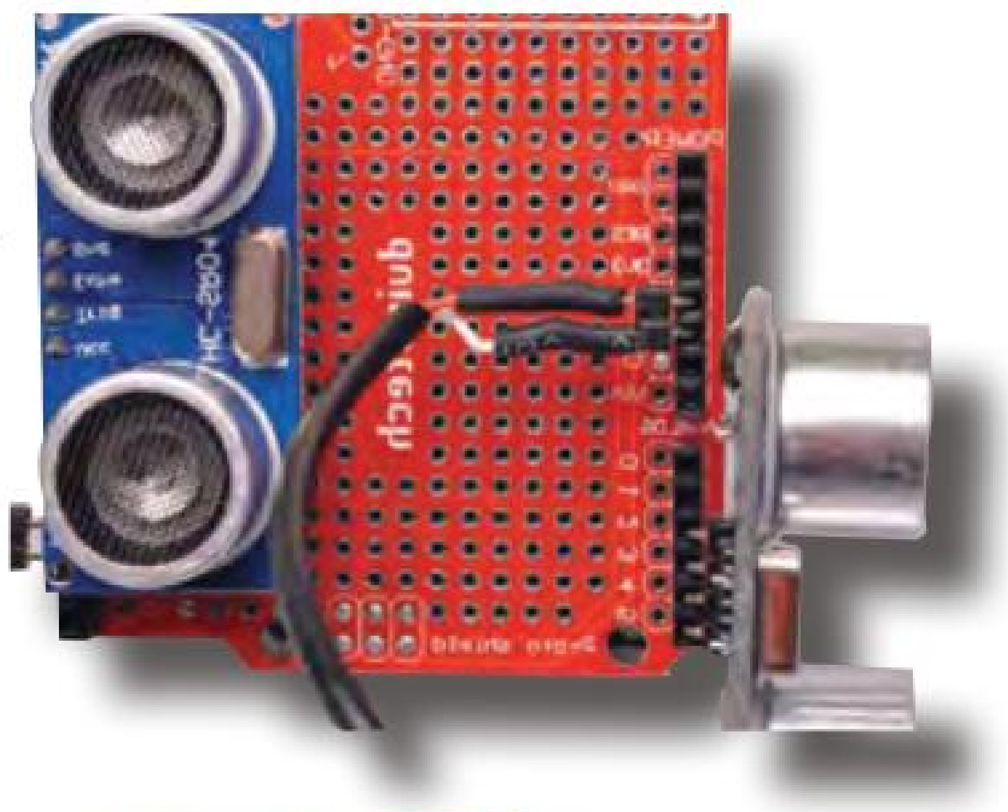


Touchscreen DDS Signal Generator

This fantastic design can produce sine, triangle or square waveforms from 1Hz to 10MHz, with ±0.005% frequency accuracy and it also has a sweep function. Its touchscreen LCD makes it very easy to drive and you can use it for audio or RF applications.

Upgrade your Arduino-based Theremin

In last December's issue we had a short article on building a simple Arduino-based digital Theremin. Next month, we'll show you how to add a second sensor to control volume.



Low Cost Electronic Modules – Part 4

Learn to use the AM2302/DHT22 digital temperature and relative humidity (RH) sensing module. It provides just about the simplest way to make a microcontroller project with temperature and RH sensing capabilities.

Teach-In 2018 – Part 7

In next month's *Teach-in 2018* we will look at radio frequency (RF) tests and measurements, and introduce a selection of RF test instruments and measurement techniques. Our practical project will feature a sensitive RF 'sniffer' that can be used to check for radiated signals over a very wide frequency range.

AM2302 DHT22 Sensor

PLUS!

All your favourite regular columns from *Audio Out* and *Circuit Surgery* to *PIC n' Mix* and *Net Work.*

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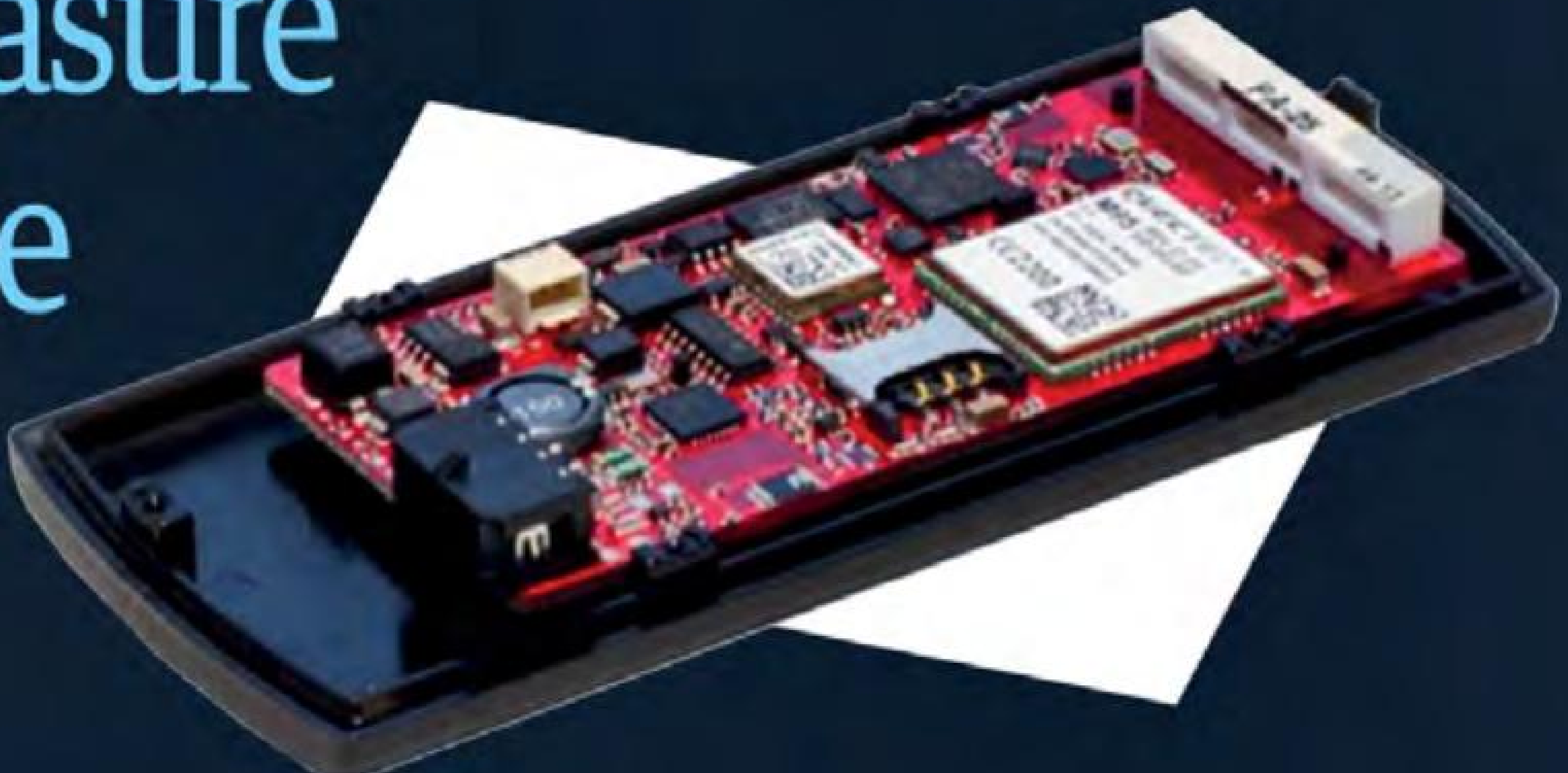

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